IBA

TECHNICAL REVIEW

23

Developments in Aerials for Broadcasting

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IBA TECHNICAL REVIEW March 1986
ISSN 0308-423X

23 Developments in Aerials for Broadcasting

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Introduction

by R. C. Hills
Assistant Director of Engineering (Operations)
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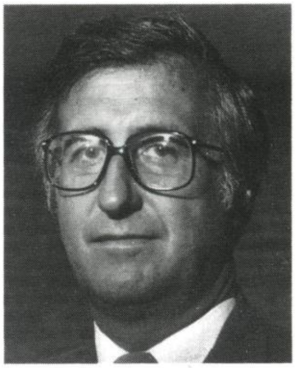


One of the most important features of any broadcast transmitting station, be it for television or radio, is the aerial system. Having raised sufficient power in the transmitter itself – which aerial engineers would always claim is the easy part of the business – the ultimate success of the station is very often determined by the way in which that power is then concentrated to deliver the strongest possible signal within the desired coverage area, at the same time wasting as little as possible in other directions. The essential features of a good aerial are that it should handle with safety the maximum transmitter power, cause minimal distortion of the transmitted signal – which is particularly important in the case of television, where a picture can be perceptibly degraded by delayed images or unwanted variations in the relative levels of the luminance and chrominance signals – and finally concentrate the power radiated in such a way, both in azimuth and elevation, that it is used to maximum advantage.

The design of radio and television transmitting aerials for use in the United Kingdom has been a process of evolution in which the broadcasters themselves have worked in close co-operation with industry to meet the successive challenges. The process of specifying what is required has to start with the broadcaster, for it is he who has the clearest idea about the area he wishes to serve and the site where the transmitter is to be built. From this starting point, and using the wealth of statistical information which has been accumulated over the years about the propagation of radio waves at various frequencies, he can derive the basic requirements for his aerial in terms of gain, height above ground and horizontal and vertical directivity. It then becomes the task of industry to translate these requirements into a practical design, developing the necessary hardware in the process. So often, and particularly in the case of aerials required to transmit several channels of UHF television simultaneously, the final result has to be compromise between the ideal needs of the broadcaster and what can actually be realised within the practical constraints of physical design. A close and continuing relationship between broadcaster and industry is essential to such an iterative design process. It is almost invariably the case that the transmitting aerial, be it for television or radio, has to be custom built to meet the requirements of coverage from any particular site.

The introduction of television services in the UHF bands and using the 625-line PAL System I specification in the late 1960's, represented one of the most significant challenges which has ever faced the aerial designer in the UK. The requirement to carry four channels simultaneously, each of high power, with minimal variation of both vertical and horizontal radiation patterns across and between the channels, and within the limitation of physical space imposed by the constraints of structural loading on existing masts, posed a series of unique problems. The combined knowledge and skill of aerial engineers within the ranks of the broadcasters and industry evolved solutions to those problems which are unrivalled elsewhere and which have been the key to the successful development of the UHF television service which now covers over 99% of the UK, with signals of high quality and reliability. Some of the articles in this issue examine the background of those designs and of the associated filter equipment needed to combine the transmitters into the aerials.

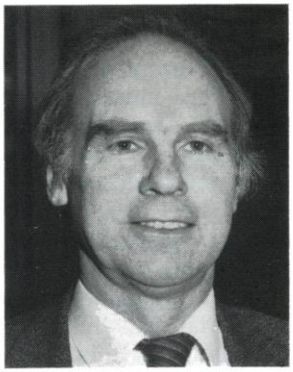
Another aspect of design of all transmitting aerials which is vital to their successful application, concerns their maintenance. Unlike the transmitters, which can readily be mended at any time they go wrong, within the environment of a warm and dry transmitter room, the associated aerials are frequently at the top of extremely high structures, always exposed to the elements, and presenting in

contrast an extremely hostile environment in which to carry out any maintenance work. Much thought has gone into the design of such aerials to ensure that, as far as possible, they can remain maintenance free and that, where preventive or corrective maintenance is required, it can be carried out as effectively as possible with minimum interruption to the service. There is no doubt, however, that the inescapable need to operate many of the component parts, particularly semi-flexible feeders, at powers very close to their maximum permissible rating, underlines the vital importance of a properly considered maintenance programme if catastrophic failure is to be avoided. The subject of maintenance will be dealt with in a later Technical Review.

The introduction of Independent Local Radio in 1972 presented new challenges to engineers of the IBA and of industry in terms of aerial design and development. Initially on MF it was necessary to make use of highly directional aerials in order to lay down sufficiently strong field strengths within the wanted service area and at the same time protect the re-use of the same frequency in other locations which could be subject both to ground-wave and sky-wave interference. A later development was the use of aerials with less directional patterns enabling the service area to be tailored to fit what was editorially desirable. On VHF, the need for circular – or at least mixed – polarisation to improve reception on portable and in-car receivers, particularly in association with a stereo service, brought other problems of design. Elegant solutions were found to all of these problems and a number of the contributions to this issue explain them in some detail.

As someone who began his career in broadcasting engineering working on the design and development of aerial and feeder systems, I am particularly pleased to have been invited to write this foreword. Those of us who have been fortunate enough to work in this area are well aware that others tend to regard the design of aerials as much an art as a science. Hopefully the articles which go together to make up this Review will shed some light upon the sound scientific and engineering bases upon which have been designed about 950 such aerial systems now in satisfactory service throughout the United Kingdom, developments in which IBA engineers have played a significant part over the last twenty years.

J. A. THOMAS, C.Eng., FIERE, joined the then ITA in 1968. In 1969 he was appointed as Head of Masts & Aerials Section and became Head of Masts & Aerials Group in 1983. Previously, he worked for the BBC starting his career in broadcasting at a transmitting station before moving to work with the Unit responsible for provision of VHF and UHF aerials.
During his period with the IBA he has also been involved in engineering consultancy tasks for developing countries in Africa.



Provision & Maintenance Of IBA Aerial Equipment

by J. A. Thomas

Synopsis
Responsibility for the provision and maintenance of all aerial equipment and structures in the IBA has been merged into a single group. A total of some 60 staff, with the majority based in the engineering regions, carry out this task. Provision of new aerial equipment is carried out by a small unit based at engineering headquarters at Crawley Court.

Aerials in use at UHF television main and relay stations are reviewed together with their power monitoring and protection equipment. Development of improvements to this equipment are discussed.

Reserve aerial and mast facilities are essential for the IBA television and radio transmission networks and a full range of interim and main reserve equipment is available in the event of catastrophic failure.

Since August 1983, responsibility for the provision and maintenance of the IBA's mast and aerial equipment has been combined in the Station Operations and Maintenance Department. The Masts and Aerials Group has some 60 staff – the majority being deployed at Regional maintenance bases. A unit based at the IBA engineering headquarters at Crawley Court is responsible for provision of new aerial equipment. The other articles in this review highlight the latest developments in television channel combining filters, television aerial impedance matching techniques and considerations for aerial support structures. Provision of aerial systems for the ILR network has also involved new developments and investigations which are described in the accompanying articles. These illustrate how UK aerial contractors have successfully met the challenge of the new requirements for the Fourth Channel transmission network (Channel 4 and S4C) and the expansion of Independent Local Radio (ILR).

Aerial Provision

Apart from a few aerials built exclusively for the Fourth Channel, and replacement aerials, the UHF television main station aerial systems were provided some years ago – most before 1969. A variety of designs has been used – some based on the use of panels of slot or dipole radiators mounted on steel poles or lattice columns surrounded by glass reinforced plastic (grp) cylindrical shrouds for weather protection. Other designs use structural grp cylinders with slot radiators supported against the inside wall of the cylinders (fig.1). The aerials were de-

Fig.1. Inside view of a UHF omnidirectional aerial installed in a 3ft diameter glass reinforced plastic cylinder. This type of aerial is used at main stations.

signed for multichannel operation, the majority radiating all four channels, (although in some cases separate aerials, each radiating two channels are used) with channel combining filters which were often installed in separate buildings near the mast base. Initially, of course, combining of only two or three channels was required and the addition of the Fourth Channel transmission facilities necessitated the provision of a two-channel combiner based on the Rotamode filter design.

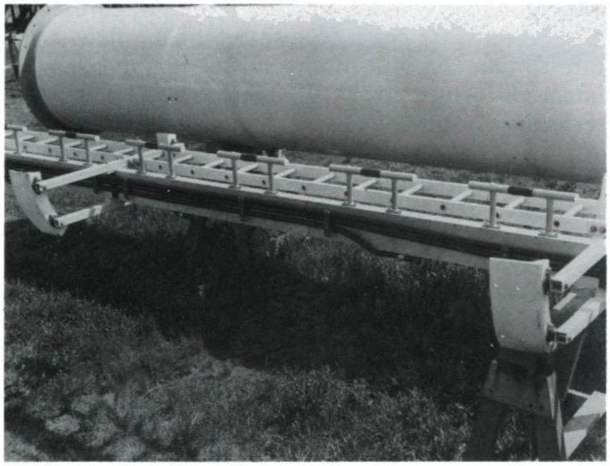


Fig.2. The dipole assembly for a vertically polarised aerial giving a cardioidal horizontal radiation pattern; the structural g.r.p cylinder is in the background. Aerials of this type are used at a number of higher power relay stations.

Earlier relay stations with cardioid horizontal radiation pattern requirements used arrays of vertical dipoles supported on an aluminium spine (fig.2) mounted inside a 3ft diameter structural grp cylinder. More directional pattern requirements were met by means of panel arrays using radiating elements printed on a glass fibre laminate. Both types of aerial have proved to be reliable in service but as the UHF network developed and in view of the smaller populations to be served it was necessary to consider a more economical solution for these low power and low cost relay stations. Several hundred low power stations with erps in the range lW to 100W are required to complete the network and the aerial arrangements employed are usually based on the use of log periodic elements (fig.3). Based on a design originally developed by the BBC Research Department to cover the whole of the UHF band, the log periodic aerial is made from aluminium. It is used as the receiving aerial as well as the basic element in the various configurations forming the transmitting array required in order to satisfy a range of

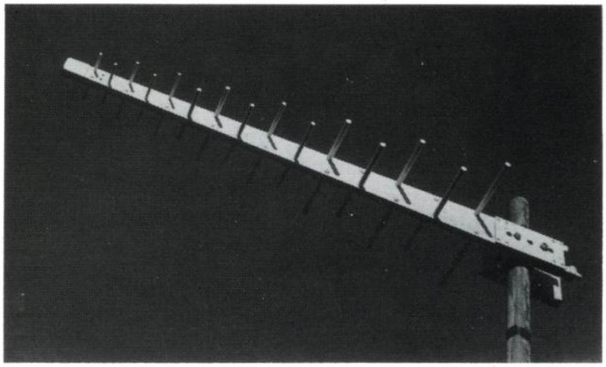


Fig.3. A log periodic aerial typical of the type used in arrays giving tailored radiation patterns at small relay stations. This aerial is frequently used as a receiving aerial.

directional pattern requirements. Mounted, typically, on a low cost 17m tripod tower fabricated in steel tubing, the aerial system is installed by the tower erectors and set to work in one to two days at site.

Low cost channel combining and splitting filters have been developed for these small relay stations using the commutating line principle in conjunction with printed circuit 3dB couplers. A complete 4-channel combining/splitting arrangement occupies only five inches of standard rack space. An IBA designed active splitter is used when the received signal is marginally low.

For main stations, aerial monitoring and protection equipment is necessary to ensure that power is removed immediately a fault condition is detected. This equipment provides overall power protection with very accurately calibrated directional couplers installed at the lower end of the main feeders. The couplers monitor the forward and reverse power which is measured by power meters equipped with thin film thermo-couple or thermistor heads. Protection against over power is also provided at most of the stations to guard against inadvertent excessive power increases from the transmitters. A limitation of the overall power measurement is that it may not detect all fault conditions in the branches of the aerial distribution feeder system since the effect of such faults is diluted.

Replacement of the earlier type of this equipment is being considered and a feasibility study is to be undertaken into a more sophisticated protection arrangement which it is hoped will overcome this limitation. Because operation of the aerial protection trip equipment results in all transmissions being off the air it must be highly reliable and not subject

to any spurious tripping. A valuable extension of the existing design would be the possibility of having a record of the aerial reflection characteristics in the period immediately before a trip occurs. It would also be helpful to have a means of determining any increases in reflections due to weather conditions.

Aerial Maintenance

The extension of the hours of broadcasting in television and with many ILR stations using 24-hour operation the difficulties of maintaining the aerial systems have increased. More and more maintenance has to be done outside normal working hours which has encouraged investment in the most efficient test equipment available based on the use of network analysers. Six aerial maintenance teams, each consisting of one engineer and one technician, are deployed from bases in the regions assisted by rigging teams who also carry out mast maintenance.

All the UHF main stations and the higher power relay stations have their aerials divided into two sections, each fed by a separate main feeder. This enables transmissions to be continued on one half aerial while the other is maintained. Work within the aerial aperture is often limited, however, by the extent of the radiated field in the climbing areas. More limitations will result if the reductions in permissible rf exposure levels which are being considered by the National Radiological Protection Board, and which are expected to be published as a UK national safety recommendation, are adopted. The IBA has developed an accurate measuring cell for calibrating rf hazard meters. The development and supply of a robust meter with improved accuracy and fitted with an audible alarm is also being investigated with manufacturers of this specialised equipment.

Full aerial overhauls are normally carried out at intervals of about 10 years with a major inspection every five years. For the older aerials, work has started on replacing some obsolescent flexible feeders which are beginning to fail, possibly due to deterioration of the dielectric. In some cases, the combination of the need for new feeders and radiating panels, together with the need for improved access, may result in complete replacement of an aerial.

Provision for maintenance, including improved access and a reduction in the level of radiation within the climbing space, will be given full consideration. Should a replacement mast be required, careful attention must be given to location of the new structure to take account of any shadowing and delayed image distortion during the period when mutual interference occurs.

Reserve Equipment

In the event of a mast collapse or catastrophic failure some means of restoring services in the quickest possible timescale is essential. For television main stations reserve equipment has been acquired for such emergencies comprising an 80m tripod tower with a cantilever omnidirectional panel aerial. This combination forms the interim reserve aerial system which would enable some restoration of transmissions within a few days. A higher level of restoration would depend on use of the main 215m reserve mast and 100 kW type aerial which would take several weeks to erect although this period would be minimised with the use of the quick-erection derrick provided with the mast. A reserve channel combiner is also available. All the equipment is ready for use at both IBA and BBC landlord stations and has been purchased jointly as a reserve facility.

For ILR, two transportable MF aerial systems are available for emergency use as well as two versatile sloping wire radiators fed by dual-frequency matching networks. Spare aerial units are also available for VHF.

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Aerial Support Structures

by M. J. Lambert

Synopsis

The author reviews the many factors which must be considered by the Broadcasting Engineer when defining a suitable support structure for a broadcasting aerial system, and he outlines current methods of design and construction of self-supporting towers and guyed masts.

The diverse range of aerial types, sizes and functions is matched by the range of structures required to carry them and the provision of any broadcasting facility involves an efficient and economical support system which often represents a significant proportion of the cost of the project.

The most spectacular structures which dominate the skyline and landscape are exemplified by the 330m concrete and steel tower (fig.1) at Emley Moor, Yorkshire or the Telecom Towers in London, Birmingham and Manchester, or even the slim pencil-like cylindrical guyed mast (fig.2) reaching 380m above the Lincolnshire Wolds at Belmont.

At the other end of the scale, many thousands of television and radio sets receive their signals from local relay transmitting aerials mounted on 20m-45m towers and 17m lattice poles (fig.3), small wooden poles and even roof-mounted brackets on tall buildings close to the service areas.

Choice of Structure Type: Mast or Tower?

The factors which determine the height and type of structure are defined primarily by the type of aerial to be used and the service area to be covered. Large aperture UHF aerials often require a long and uniform structural cross-section of minimal face width on which dipoles or panels are directly mounted and this structural constraint generally means that a guyed mast is most suitable. Conversely the requirements of SHF link dishes demand larger face-widths for mounting and a high degree of resistance to

Fig.1. The 330m concrete and steel tower at Emley Moor, Yorkshire.

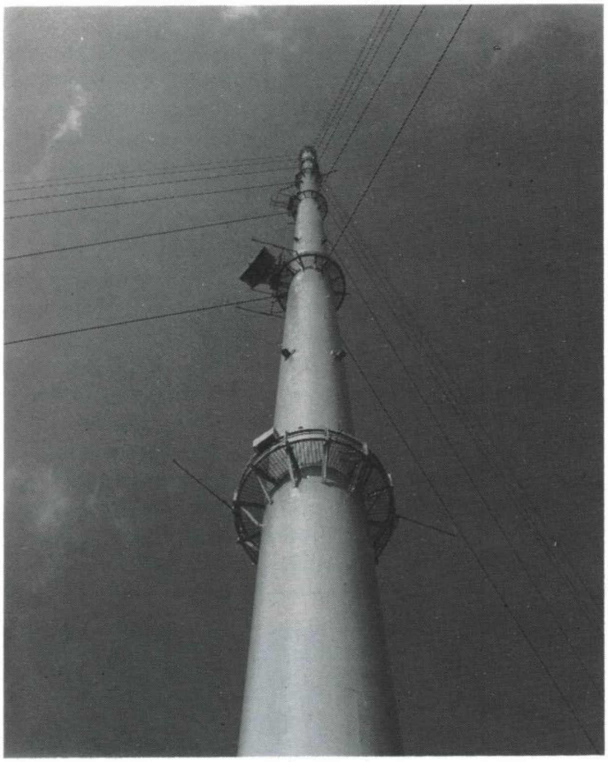


Fig.2. The 380m tubular steel guyed mast at Belmont, Lincolnshire.

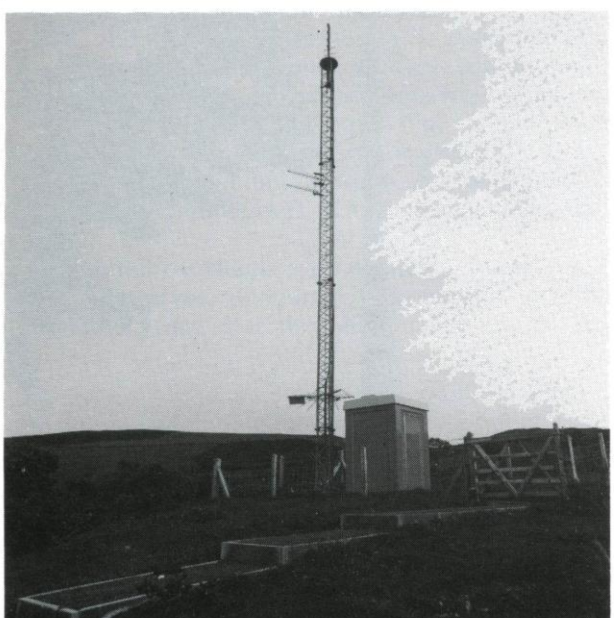


Fig.3. A typical 17m lattice tower as used at many small relay stations.

angular structural deflection under extreme wind loading, so heavier construction self-supporting towers are usually more suitable (fig.4).

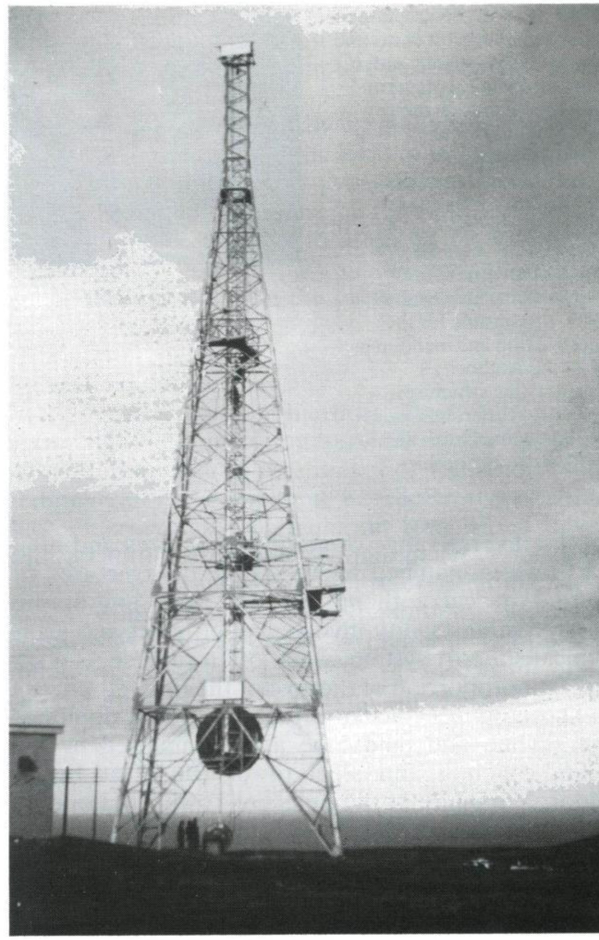


Fig.4. The self-supporting tower used at the Fair Isle SHF link station. High wind speeds and the loading imposed by the SHF dishes require the use of this type of structure.

Compromises are frequently necessary. A single structure is often required to carry UHF omnidirectional aerials, VHF radio, SHF link dishes and sundry monitoring and radio-telephone arrays and the structural designer must then consider the following factors:

(i) Required mean aerial height for each aerial system,
(ii) Wind loading on each element of the array,
(iii) Size, weight and disposition of all feeders,
(iv) Ice formation on steelwork and aerials,
(v) The permitted angular deflections in azimuth

and elevation of each aerial above which the broadcast signal is significantly affected,

(vi) The need for all-weather access to some or all of the aerials,

(vii) The expected peak wind loading during the life of the structure,

(viii) The degree of security required,

(ix) The available ground area and access to the site,

(x) The geological nature of the site,

(xi) The overall cost of land, foundations and structure,

(xii) The cost and implications of future maintenance or of structural replacement,

(xiii) Any special planning considerations imposed by statutory bodies,

(xiv) The aesthetic appearance of the structure.

Any one of these factors can influence, or even define, the primary choice between a free-standing tower and a guyed mast. The structural weight (and therefore the cost) of a typical lattice tower varies approximately as the square of the height (H) whereas that of a guyed mast varies approximately as $H^{1.5}$.

In the United Kingdom, where land prices are comparatively high, the 'break point' has remained fairly constant over the past thirty years at around 100m, above which towers become uneconomical.

The main exception to this generalisation is the MF radiator where the structure itself forms the radiating element. The cost of insulating all the legs of a tower outweighs that of a single mast base insulator and associated stay insulators. Moreover, the form of the mast body is itself generally electrically superior in radiating quality and such installations usually incorporate one or more guyed masts, unless considerations such as the need for (and high cost of) piled foundations, or a very restricted site area makes a tower more suitable.

Loading and Design

The principal loads applied to any structure arise from the deadweight of the aerials, feeders and associated equipment, and the accretion of ice on these and on the steelwork. Dynamic wind loading is imposed by the aerodynamic drag on the structure, aerials and accumulated ice and this loading is a function of wind speed, height above ground level, nature of surrounding terrain and the shape and roughness of each wind-loaded item. Many full scale experiments and wind-tunnel tests on models have been carried out throughout the world to determine empirical rules for assessing these loads and much of the data has been incorporated into national design codes.

Initial design analysis assumes a steady-state 'quasi-static' loading arising from the maximum design wind force being applied continuously to the structure to enable the primary stresses (and thus the necessary size and strength of each component) to be calculated assuming a suitable factor of safety for each.

Once the final shape and size of every component has been so determined, a dynamic analysis check is made, taking into account the effect of possible wind gusts, 'patch loading' (i.e. non-uniform wind profiles), the effects of aeroelastic excitation and of mass-distribution and any structural damping, and this often highlights areas where increased strength is required.

In the UK the British Standards Institution Code of Practice CP3 has, until recently, been the design standard used but because this code was formulated primarily for buildings, a new code has recently been developed (BS CP Lattice Masts and Towers – Part 1, Loading) which enhances and supersedes CP3.

More recent design methods have changed from the earlier 'permitted stress' concept incorporating global safety factors to a 'limit-state' philosophy whereby each element of each loading condition is modified by a series of variables known as partial factors, which reflect the degree of confidence in the calculation of the predicted loads and the statistical chances of these being exceeded. When the variations of all loading effects which could occur simultaneously on a component are then summed, the component itself can be designed to withstand this maximum, again using predicted strength values amended by partial factors to take into account the quality of workmanship, the source of materials and the predicted future maintenance level.

An assessment of the design wind speed is made for any site from interpolated meteorological records and is based on the extreme value of the mean hourly wind speed occurring once in 50 years at a height of 10 metres above ground level. Isopleth maps of the UK (fig.5) and other land masses (e.g. USA, Australia, Northern Europe) are readily available and the basic windspeed (V_{10}) derived from these is then adjusted by partial factors to take account of height, prevailing wind direction, nature of terrain (which influences the boundary layer and thus the wind gradient), and the 'serviceability cri-

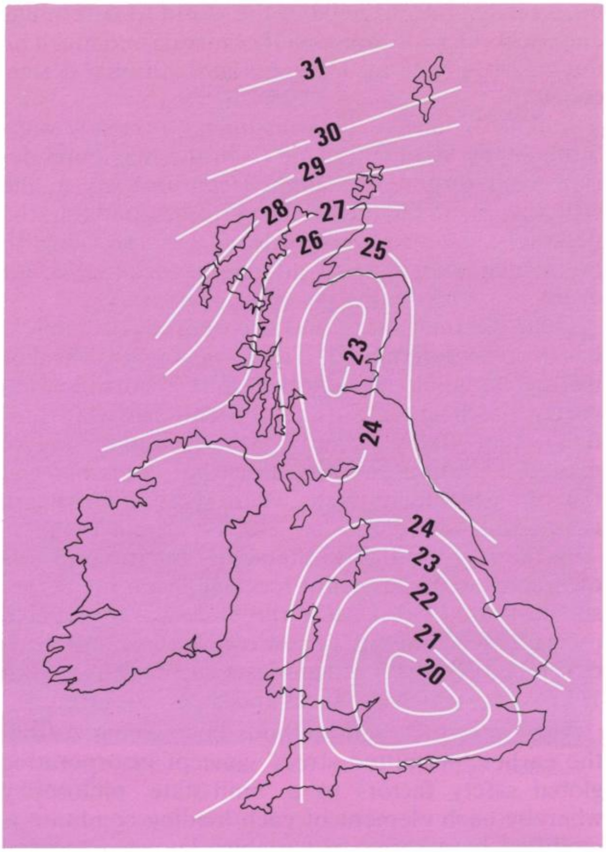


Fig.5. Basic mean hourly wind speeds, V_Bm/s, (at sea level in open terrain) for the United Kingdom.

teria' or degree of security defined elsewhere in the code ranging from 'temporary installation' to 'Communication structures vital to Defence or National Security'. The pressures thus calculated are applied to the effective wind areas, taking experimentally determined drag factors into account, and give the imposed design loading for limit-state analysis. A further new 'limit-state' British Standard for stress analysis and detailed design of structural steelwork, introduced in 1984, supersedes BS449 which has been widely used for many years in the design of masts and towers, while 'Eurocode No. 3 – Steelwork' is being formulated as an international strength code for use throughout EEC countries.

Construction

The majority of tall structures are of lattice construction comprising either rolled steel angles, tubes or a combination of both. For ease of erection, joints are generally bolted to enable each member or sub-assembly to be separately galvanised, transported to site and winched up into position on the structure thus obviating the need for on-site welding or riveting. Smaller towers, up to approximately 25m high, are often assembled horizontally and then lifted onto prepared foundations by crane, but where crane access is difficult, and for taller structures, a climbing derrick and ground-mounted winches are required.

It is important that a guyed mast is kept truly vertical thoughout the erection procedure, using its permanent guys and intermediate temporary guys, otherwise permanent distortions can be built into the mast column which make final vertical adjustments very difficult to achieve.

Lightning protection, earthing and aircraft warning lighting must be progressively installed as erection proceeds to minimise risk; and ladders, platforms and hand-railings complying with safety legislation form a significant part of the ancilliary steelwork subsequently installed.

At all times the safety of personnel working both on the structure and on the ground below is paramount and this frequently precludes many operations when the windspeed exceeds 10-15 knots, when the steelwork is icy or wet, or when lightning is imminent. These constraints make the detailed programming of construction extremely difficult to effect with accuracy.

Special Structures

When a structure is located close to public roads or housing, the disposition of stay ropes for a tall mast can be an insuperable problem and a free-standing tower, albeit more expensive, is required. Such is the case at Crystal Palace and at Croydon where 230m and 155m towers serve virtually the entire Greater-London area with UHF TV, VHF radio and SHF links. At Emley Moor in Yorkshire, a 274m concrete tower (fig.1) supports a 56m steel extension on which is installed a four-channel television aerial within a glass-reinforced plastic (grp) cylinder. This form of construction has the advantage of all-weather access to the aerials, and internal housing within the base area for much of the SHF link equipment. Outside-Broadcast link facilities are available from an enclosed equipment room forming a 'turret' at the top of the concrete section.

All-weather access is also afforded by the cylindrical shell construction of the IBA's 380m guyed mast (fig.2) at Belmont and a similar mast at Winter Hill, Lancashire. These comprise a 2.75m diameter hol-

low steel column surmounted by a grp-shrouded steel lattice spine. The primary disadvantage of any such circular shaped structure is a tendency to aerodynamic oscillation at moderate windspeeds which can cause large and fluctuating aerial displacements and high structural stresses unless measures are taken to attenuate the amplitude of oscillation.

The adverse consequences of large amplitude oscillations are not only cyclic variations in the level of broadcast signals in some directions, but also fatigue loading of structural components which causes a reduction in their serviceable life. Close attention to detailing and quality control of welding and fabrication, together with the choice of suitable materials will minimize but cannot eliminate this effect.

Sensitivity to vortex-induced excitation can be reduced by fitting helical 'strakes' or some other type of aerodynamic spoiler to disrupt the airflow around the cylinder, but the resulting increase in drag generally imposes greatly increased wind loading at extreme windspeeds for which allowance must be made in the initial design. Mass-damping devices (e.g. heavy steel chains suspended in butyl-rubber lined tubes) tuned to the natural frequencies of highly sensitive masts have been successfully and effectively installed.

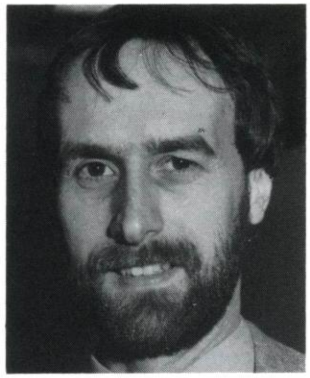
Maintenance

Access to broadcasting structures for maintenance and inspection is limited by weather conditions and by the constraints of the hours of broadcasting where powered aerials could constitute a hazard to personnel. Over the past decade the extension of broadcasting services and hours has resulted in a large proportion of structural and aerial maintenance and inspection being undertaken at night. This has led to the development of techniques for working with floodlights and personal lighting on tall structures. However the design of any structure and aerial system needs to take account of the difficulties of such working and to eliminate, where possible, any details which could create a maintenance problem. Regular and preventive maintenance (painting, stay greasing, bolt checking etc) ensures that the structure remains fully serviceable for its design life and should result in a very long life in situations where motion does not result in significant structural fatigue.

The most effective protection for exposed steelwork is a hot-dip zinc galvanised coating (complying with BS729) which can have a life of up to thirty years in a rural environment. Regular painting of structural steel, at intervals ranging from two to seven years, is necessary in coastal or industrial areas depending on the degree of pollution.

Ancilliary items such as grp weathershields and aerial cylinders also need regular inspection and maintenance. Experience has shown that the serviceable life of grp is generally less than that of its steel support (due to fatigue, ultra violet degradation, water ingress and chemical deterioration). The need for possible total grp replacement is a matter which should receive consideration at the initial design stage.

NIGEL PORTER, BSc, PhD, MIEE, C.Eng, graduated in 1976 with a first class honours degree and diploma in Electronics and Systems Engineering from Hull University. He spent 3½ years at the NRAL at Jodrell Bank on research in the field of Radio Astronomy, obtaining a PhD in July 1981. He joined the IBA in November 1981 and since November 1982 has worked as an aerial engineer in the Masts and Aerials Group, SO&M.



An Investigation Into Aerial Pattern Distortion By Re-radiating Structures

By N. D. Porter

Synopsis
The numerical moments method of analysis can be used to analyse the radiation patterns of simple antennas in the presence of re-radiating structures. Under such conditions an assumption of uniform or sinusoidal current distribution is inadequate and a solution must be found for the true current distribution on both the antenna and its support structure. The VHF Band II 'Loop-Dipole' array supported on a cantilevered pole is analysed in detail using this method with software written for use on the IBA's Honeywell computer.

INTRODUCTION

Circular polarisation can be achieved by a combination of horizontal and vertical antennas provided that the electric fields produced are equal in magnitude and in time phase quadrature. If the individual antennas produce suitable horizontal E_H and vertical E_V electric field radiation patterns then in theory it is possible to obtain a nearly omnidirectional circularly polarised pattern E_T. Indeed any E_T pattern may be constructed.

A simple practical application of this combination is a vertical dipole and a horizontal circular dipole usually referred to as a loop dipole. The vertical dipole generates an omnidirectional E_V pattern in the horizontal plane. To generate a truly omnidirectional E_H pattern an ideal constant current loop would be needed. However, in practice a circular dipole is used, consisting of a half wave dipole bent into circular form with capacitor plates attached to each end. The capacitance across the gap prevents the current from falling to zero at the ends and a very nearly uniform radiation pattern can be achieved. In practice, depending on the value of this capacitance, the current at the gap may fall to about 0.2 I_{max} resulting in a 'sausage' shaped E_H pattern (fig.1).

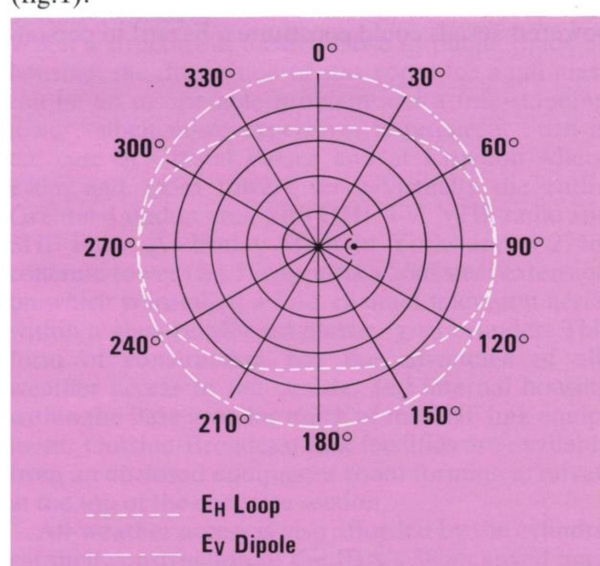


Fig.1. Computed radiation patterns for a theoretical loop dipole array.

The C & S Antennas' Band II 'Loop-Dipole' array, shown in fig.2, is a commercial application of

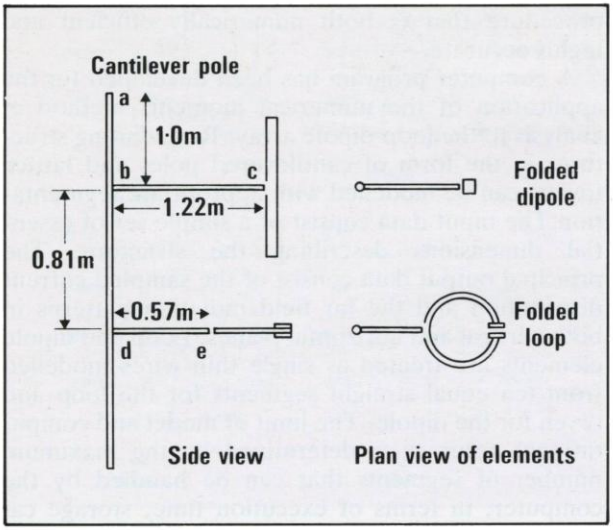


Fig.2. Basic diagram of the C & S Antennas' Loop Dipole circularly polarised array as used at IBA VHF local radio stations.

this principle and has been extensively used by the IBA for low power VHF ILR systems. In practice, however, the radiation patterns predicted in free space are modified by the presence of the support structure. The resulting patterns are unpredictable and often the aerial may require physical adjustment at both factory and site to ensure the pattern fits the desired template. Even in the simple pole-mounted case typical E_H patterns show distinct asymmetries. It had been suspected that re-radiation from the antenna booms might be the main cause so a computer modelling investigation was carried out using moments methods to gain an insight into the causes of the asymmetries and how they may be suppressed.

Loop Dipole Theory

Consider a horizontal loop with a vertical half-wave dipole piercing the centre. The far field equations are[3]:

$$E_V = j\frac{k_1}{r} I_D e^{-j2\pi r/\lambda} \quad \ldots \text{(i)}$$

$$E_H = -\frac{k_2}{r} I_L e^{-j2\pi r/\lambda} \quad \ldots \text{(ii)}$$

where:

I_D	= dipole current
I_L	= loop current
k_1, k_2	= constants for a given loop and dipole
r	= distance from centre of loop to far field point

From (i) and (ii) the vertical and horizontal fields are in phase quadrature when the currents in the loop and dipole are in phase. Circular polarisation can therefore be achieved if the magnitudes of E_V and E_H are made equal by ensuring the correct ratio of I_D to I_L.

The quadrature phase relationship can be demonstrated by integrating the current distribution on the loop, assuming for simplicity a constant loop current, and using the centre of the loop as the phase reference.

In practice, exact coincidence of phase centres may not be possible, but in principle circular polarisation can be restored by suitable phasing. It should be noted that since constant loop current cannot be easily achieved in large loops then it is not possible to achieve exact circular polarisation at all points on the pattern.

Moments Methods

In principle there is no great difficulty in determining the radiation field produced by an arbitrary current distribution. The difficulty arises when the current distribution is not known and may be a complex function of a localised source current and an induced current impressed, for example, by a re-radiating structure. The main obstacle then is to deduce the true current distribution. Under these conditions on arbitrarily shaped structures the most rigorous method is based on a numerical moment method of analysis[1].

To obtain accurate solutions for wire antennas the current distribution on the wire must be solved, subject to the boundary condition that the tangential electric field is zero along the surface of the wire. In general this approach gives rise to an integral equation of the form:

$$\int I(z')\, F(z', z)\, dz' = E_s(z) \quad \ldots\ldots\ldots\ldots \text{(i)}$$

for a z-directed current element. $E_S(z)$ is the field radiated in free space by the equivalent current $I(z')$. A solution for $I(z')$ must be found. The method of moments is essentially a procedure for reducing this

integral equation to a series of linear simultaneous equations in the generalised form:

$$[Z_{mn}]\ [I_n]=[V_m] \text{ for } n,\ m=1,\ \ldots\ N \quad \text{(ii)}$$

which can be solved by numerical methods.

The unknown currents $[I_n]$ are a suitable expansion of $I(z')$. The radiating structure is replaced by a thin wire approximation and divided into N small segments. The integral equation (i) is enforced at each segment m, subject to the boundary condition:

$$E_s(z) = -E_i(z) \quad \text{(iii)}$$

where $E_i(z)$ is the impressed field at segment m, and is known from the driving sources located on the antenna. Hence the elements of $[V_m]$ are obtained.

The reduction of equation (i) to (ii) is achieved by replacing $I(z')$ with a piecewise sinusoidal expansion of the form:

$$I(z') = \sum_{n=1}^{N} I_n\ K(z) \quad \text{(iv)}$$

where I_n is a complex expansion coefficient and $K(z)$ is a sinusoidal expansion function. $F(z', z)$ is then written in discrete form as a generalised impedance matrix $[Z_{m,n}]$ which describes the electromagnetic interactions between segments. The elements of $[Z_{m,n}]$ are computed by classical induced EMF theory invoked by enforcing reaction tests of equation (ii) with a sinusoidal testing function[2], terms of $[Z_{m,n}]$ being the self and mutual impedances between elements. Standard matrix algorithms applied to equation (ii) are then used to obtain the coefficients of $[I_n]$.

Thus the current distribution $I(z')$ is obtained in discrete form in terms of the coefficients $[I_n]$ of the chosen function at the junctions of the segments. Along the segments the current distribution can be approximated by the sinusoidal expansion functions themselves, and the radiation pattern obtained from the integral equation using the computed current distribution. In contrast to normal field calculations which assume either a uniform or sinusoidal current distribution over the entire antenna, the moments method is a distinct improvement. An accurate representation of the true current distribution on each segment is modelled as an appropriate piecewise sinusoidal expansion which closely fits the discrete current distribution calculated from equation (ii). In this way the total far-field radiation pattern can be obtained from the summation of the fields generated from each sinusoidal term in the expansion appropriate to each segment in turn. Use of piecewise sinusoidal expansion and testing functions leads to a procedure that is both numerically efficient and highly accurate.

A computer program has been developed for the application of the numerical moments method of analysis to the loop-dipole array. Re-radiating structures in the form of cantilevered poles and lattice towers can be modelled with appropriate segmentation. The input data consist of a simple set of essential dimensions describing the structure. The principal output data consist of the sampled current distribution and the far field radiation patterns in both vertical and horizontal planes. Loop and dipole elements are treated as single thin wires modelled from ten equal straight segments for the loop and seven for the dipole. The limit of model and computational accuracy is determined by the maximum number of segments that can be handled by the computer, in terms of execution time, storage capacity and rounding errors.

Model Results

As a test case the configuration illustrated in fig.1 has been modelled, this is a single tier on a cantilevered pole.

(i) *Loop only driven*: the computed E_H pattern (fig.3a) is virtually identical to the free space pattern. There are no pole or boom currents and hence no E_V field.

(ii) *Dipole only driven*: the computed E_V pattern (fig.3a) is a predictable modification of the free space pattern and is a result of induced currents on the pole. The peak in the pole current distribution is 5% of the maximum dipole current. However there are also significant currents flowing on both booms, with a 5% peak in the boom current distributions. An E_H field is therefore produced, the pattern being shown in fig.3b. The maximum current induced on the loop is less than 2% of the maximum dipole current.

(iii) *Loop and Dipole driven*: the E_V pattern is virtually identical to case (ii) but the E_H pattern now displays distinct asymmetry with a small null at $-12°$ (fig.3c). In comparison with the pattern measured in practice (fig.3d) both E_V and E_H are in close agreement. The current distribution on the loop remains symmetrical to within 2% of case (i) and the phase is constant to within 3°. The asymmetry is not therefore

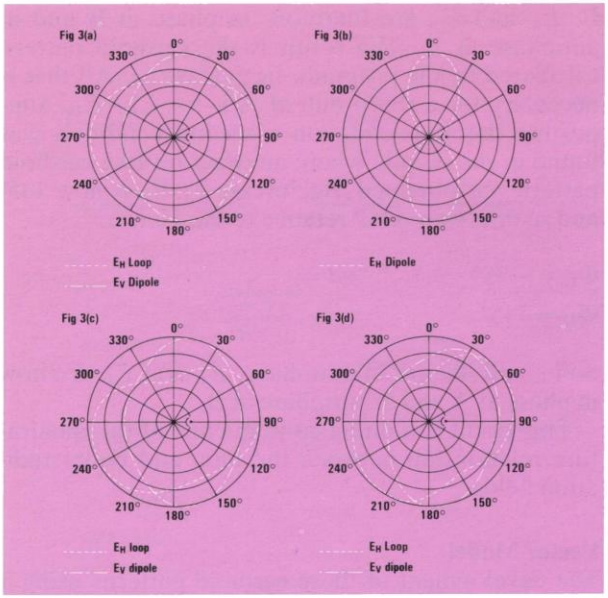


Fig.3. Computed radiation patterns for a single tier loop dipole array
(a) Loop only driven and dipole only driven.
(b) E_H pattern due to the induced boom currents when the dipole only is driven. The outer limit corresponds to 7% of that of fig.3a.
(c) Loop and dipole driven, computed pattern.
(d) Loop and dipole driven, measured pattern.

caused by a displacement in the loop current distribution.

Comparing case (iii) with (i) and (ii) it is clear that the distortion of the E_H pattern is due to the interaction of the original E_H loop field and the additional E_H field generated by the boom currents.

Boom Currents

It is only the vertical dipole that is responsible for the induced boom currents. In theory, however, with either an infinitely long pole or a finite pole symmetrically placed parallel to the axis of the dipole, there should be no induced current on the dipole boom. It seems that the physical limitation imposed, in practice, by mounting the dipole close to the top of a short pole may introduce the asymmetric loading which generates the unbalanced boom currents. In addition there will be weak coupling directly between the dipole and both the loop and loop boom, in this case, as they are also asymmetrically placed relative to the dipole axis. Also some portion of the pole current must be diverted onto the loop boom. The latter would seem to be the dominant mechanism.

The induced boom currents can be reduced by minimising the coupling between the dipole and the support structure. A study of the relationship between the computed boom currents and the relative positions and lengths of the booms indicates that, in general, the coupling can be minimised by suitable choice of support geometry.

In the sample of models tested the correlation between pattern asymmetry and boom currents is conclusive. There are five independent parameters within a fixed pole aperture so only a small sample illustrating the maximum and minimum pattern distortions are shown in fig.4. Pattern c is particularly

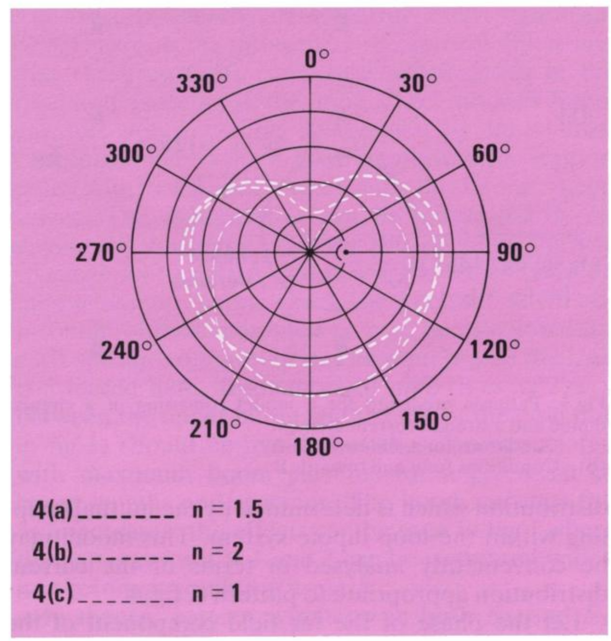


Fig.4. E_B patterns for differing support geometry such that dimension a-c (fig.1) = n./2
(a) n = 1.5
(b) n = 2.0
(c) n = 1.0

interesting in that almost complete cancellation occurs on bearing 0°. The associated boom currents show a remarkable relationship both with each other and the loop current, being in phase with each other and at +90° relative to the loop current. In this case the length a-c in fig.1 is an exact half wavelength. This is an ideal opportunity to develop a simple model for the resultant radiation field.

A Model for the Asymmetrical Patterns

The interaction between the boom and loop radiation fields is complex but a good approximation can be predicted using a simple model. The model con-

sists of a circular dipole and a single straight current element, the resultant of the two boom currents (fig.5). The resultant boom current has an arbitrary

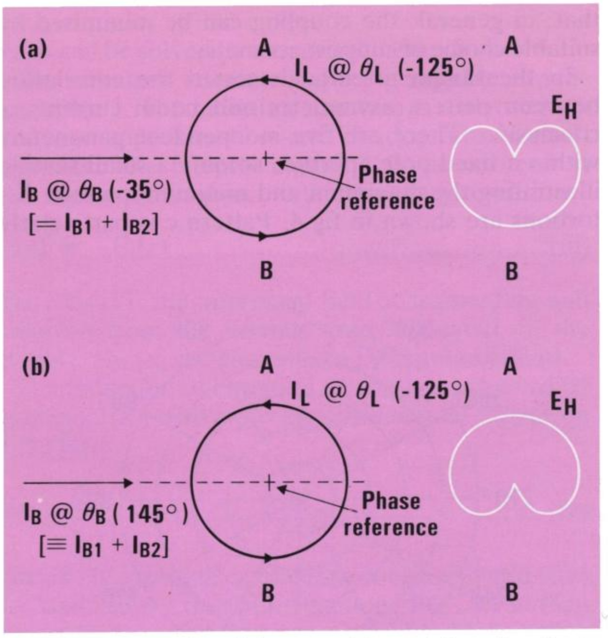


Fig.5. Patterns predicted for a model consisting of a circular dipole and a straight current element
(a) Conditions for a null towards A
(b) Conditions for a null towards B.

distribution which is determined by the mutual coupling within the loop dipole system. This model may be conveniently analysed in terms of the current distribution appropriate to pattern c, fig.4.

Let the phase of the far field component of the loop field E_L be Φ_L and boom field E_B be Φ_B. The loop and boom current phases are θ_L and θ_B respectively.

From the general field equations the phase Φ_L is +90° relative to a current element at the phase centre of the loop. For observers at A and B the phase centre of the loop coincides with the boom. Φ_L is therefore +90° relative to Φ_B for equal θ_L and θ_B. In this case, however, θ_L is −90° relative to θ_B, hence:

$$\Phi_L = \theta_L + 90° = -125° + 90° = -35°$$

$$\Phi_B = \theta_B = -35°$$

Φ_L and Φ_B are therefore in phase. However, the direction of current flow I_B, relative to anticlockwise loop current, reverses between observers at A and B. E_L and E_B are therefore in phase at B and in antiphase at A. The result is the cardioid pattern calculated by the program, fig 5a(case 1). All that is necessary for a sharp null at A is for $E_L = E_B$. Supporting this interpretation a second condition was found in the sample where an exact reverse cardioid pattern was obtained, fig.5b(case 2). θ_B is now 145° and so θ_L is now +90° relative to θ_B. Hence:

$$\Phi_L = -35°$$

$$\Phi_B = 145°$$

So Φ_L and Φ_B are in antiphase. E_L and E_B are now in phase at A and in antiphase at B.

This result is a direct consequence of the quadrature relationship between the loop and boom radiation fields.

Vector Model

The development of these cardioid patterns using a simplified vector analysis of the computed loop and boom currents is shown in fig.6. For an observer at A, the radiation field from the loop may be considered to originate from an equivalent current element I_R derived by vector addition of the contributions from all the elements of the loop at A. To a good approximation I_R is dominated by the diametrically opposite broadside elements I_1 and I_2. This can be shown with a simple integration of the constant current in an ideal loop. By restricting the integration to the two diametrically opposite arc segments corresponding to I_1 and I_2 the far field falls to only 70% of the full loop field.

Considering case 1 (fig.6e), although the phases of the resultant loop and boom vectors I_R and I'_B are both −65°, the reference direction differs by 180°. For an observer at A the two radiation fields are therefore in antiphase and cancel. Similarly for an observer at B the reference direction for I_B' has switched through 180° relative to I_R. The two radiation fields are now in phase at B and add. Hence the resulting pattern is a cardioid and the depth of the null depends on the relative magnitudes of I_R and I_B'. In this case almost complete cancellation occurs, so I_R and I_B must be virtually identical.

In case 2, the boom current phases are again identical, ie 145°, but now 180° different from case 1. Applying a similar analysis the null now occurs for an observer at B, hence the reversed cardioid pattern relative to case 1.

The computed pattern of fig.4c can therefore be demonstrated using a simple vector model. This ap-

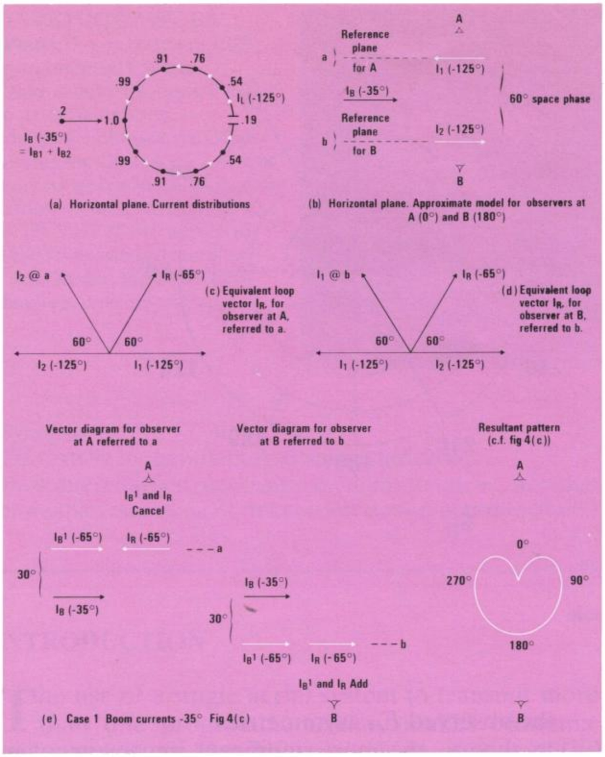


Fig.6. Derivation of the cardioid pattern by vector analysis.
(a) Assumed distribution of currents in the horizontal plane.
(b) Equivalent current distributions in the horizontal plane as seen by observers at A and B.
(c) Vector diagram for loop currents as seen by an observer at A with a as reference.
(d) Vector diagram for loop currents as seen by an observer at B with b as reference.
(e) Case 1 conditions of fig.5a (see text) giving rise to the pattern of fig.4c.

proximate analysis works well under these unique conditions. In general, however, the loop and boom current distributions bear no distinct phase relationship to either each other or the loop, and the patterns are difficult to predict. An exact analysis must take into account the full current distribution on the loop. The computer program does this, and computes the radiation field expected from any set of current distributions. From the sample modelled the predicted patterns display similar asymmetries to those observed.

Model Observations

(i) For any other phase relationship between I_B and I_L maximum cancellation will occur on a different bearing and the depth of the null will depend on the magnitude and phase of the induced currents.

(ii) The pattern nulls are confined to a ±30° range around bearings 0° and 180° as a consequence of both the inherent E_H pattern, and the increasing contribution from the boom currents as the booms themselves appear broadside to the observer.

(iii) In the specific case considered in fig.3 the computed and observed nulls on bearing −12° and −14° respectively correlate closely with a −15° change in I_B relative to case.1. (fig.5a and 6e).

Pattern Improvement

The results clearly demonstrate that significant boom currents are induced by the vertical dipole and that these currents can cause asymmetries in the radiation pattern of the loop. The induced boom currents appear to be determined by the mutual coupling between the vertical dipole and the support pole, and to a lesser extent with the booms themselves. The asymmetries can be minimised by reducing the coupling, however this demands vertical spacings and relative boom lengths which may be impractical in many cases. In general, such adjustment of these parameters must be calculated for each specific configuration. For example, in the case considered here, resonances of $n\lambda/2$ ($n = 1, 2, \ldots$) between the upper boom and the top of the pole (a-c in fig.1) should be avoided as these are associated with maximum boom currents for a given set of boom lengths and spacings. The boom currents fall as n increases; the effects can be seen in fig.4 where the patterns for n = 1 and 2 can be compared with a non-resonant condition n = 1.5.

In theory, pattern symmetry in both E_V and E_H can be restored by preventing pole currents flowing. This may be achieved by decoupling the pole from the dipole. Improvement in E_H can be affected by decoupling the booms from the pole. Neither appears readily practical, but the latter can be simulated in the computer program.

Decoupled Booms

To test this theory the computer model was run with both booms insulated from the pole. Fig.7a illustrates the effect on the pattern of fig.3c. Calculated boom currents are only 1% and the asymmetry disappears.

Fig.7b illustrates the effect on the pattern of fig.4c. The cardioid pattern disappears but some asymmetry is still present. With only the dipole fed, the highest boom current is 3.5% (compared to 18% when the booms are uninsulated). Clearly the most

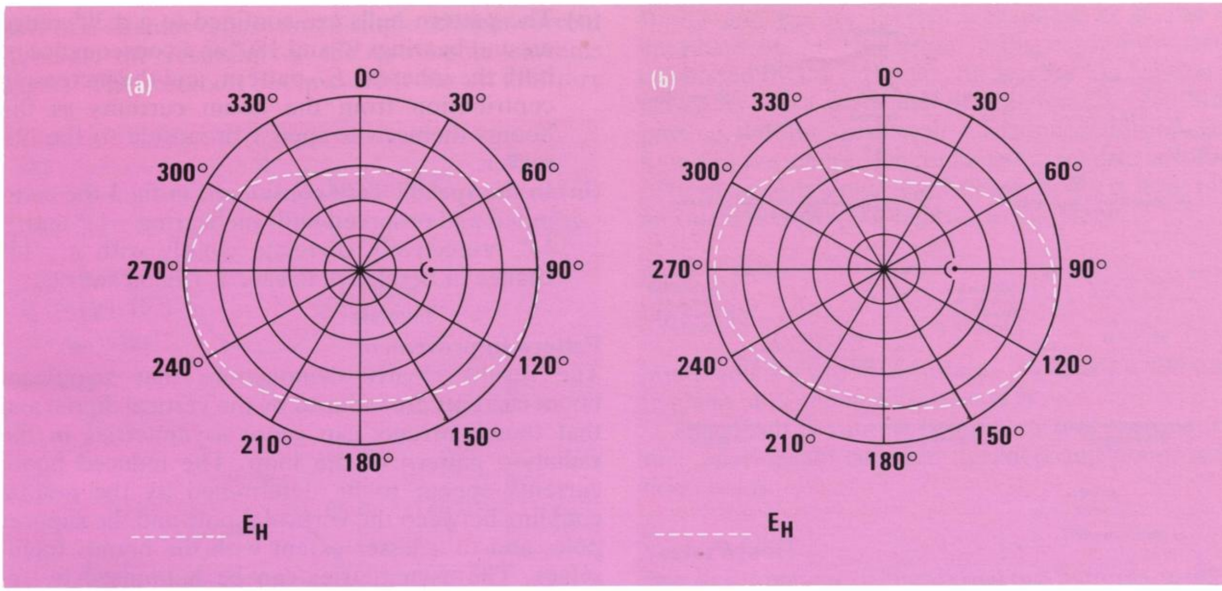


Fig.7. Computed E_H patterns with booms insulated from the support pole
(a) For the conditions obtaining in fig.3c.
(b) For the conditions obtaining in fig.4c.

significant boom currents in this case flow directly from the pole rather than by direct coupling between the dipole and the booms. The residual asymmetry represents the contribution of the direct coupling, this is still significant in this case and of the same order as seen in the configuration of fig.3c.

Polarisation

The modification of both the E_V and E_H fields by induced currents on the support structure will alter their relative phase centres. Polarisation will not remain circular and is difficult to predict. In practice, once a suitable mechanical arrangement has been found to equalise E_V and E_H, circular polarisation can be restored by measuring the phase of the far field E_V and E_H, and adjusting the phase of the loop or dipole feeds accordingly. Correct alignment of these arrays is therefore far from simple and often a compromise.

Conclusions

Despite the limitations of the modelling technique, conclusive results have been obtained and the following observations may be made:

(i) Significant boom currents are induced by the vertical dipole via interaction with the supporting pole.

(ii) These boom currents are solely responsible for the observed E_H asymmetries.

(iii) In theory, the most significant improvement in pattern symmetry can be affected by decoupling either the booms from the pole or the pole from the dipole.

(iv) The computed model patterns are in good agreement with observed patterns.

(v) The program has the potential to analyse and improve the patterns of existing arrays.

As a result of this study a versatile computer program based on the numerical moments method has been developed. An analysis of this nature would not be practicable by any other means. The program can also be used for the more general analysis of simple VHF dipole arrays mounted on similar structures and has been used to verify both the predicted and measured patterns produced by the C & S Antennas' omnidirectional 'slant dipole array'. The program has also been used to assess the implications of the Band II frequency changes on existing loop dipole patterns.

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Dual-Channel MF Aerials

by D. Cowans

Synopsis
This article looks into the particular difficulties encountered when powering a medium-frequency aerial at more than one channel. It concentrates on non-directional systems and demonstrates the ways in which optimum performance can be obtained, including a case study. Filter performances are dealt with in some detail.

INTRODUCTION

The use of a single aerial system to transmit more than one programme usually makes economic sense on any wavelength, particularly so at medium-wave where the cost of the actual aerial – a tall mast or tower – often forms a substantial part of the overall cost of the transmitting station. These economic restraints, together with the understandable reluctance of Planning Authorities to allow several single-channel stations to be built when one multi-channel station would suffice, have meant that dual or multi-channel aerials have been used by both the IBA and the BBC whenever practicable.

The first dual-channel MF aerials commissioned by the IBA were the temporary and permanent aerials installed to transmit LBC and Capital Radio in London in the early Seventies – (at Lots Road and Saffron Green respectively). The design for the latter system was produced by North American consultants because the IBA had no practical experience at that time in the design of relatively high-powered (40kW total) four-mast directional aerial systems such as that required at Saffron Green.

Since 1973 the IBA has installed many single-channel MF aerials for the transmission of ILR but only in the last few years have further dual-channel aerials been commissioned by the IBA – allowing the BBC to share aerials at some existing and some new ILR stations. The aerials have all been single-element omnidirectional radiators with the exception of the 4-mast array at Langley Mill, Birmingham, which was converted to two-channel operation in 1981.

Types of MF Radiator

Medium-wave radiators generally comprise some form of monopole fed against a groundplane. The presence of a conducting groundplane (produced by the ground itself and normally augmented by radial copper wires buried just below its surface) enables the monopole to produce similar radiation patterns, in both the horizontal and vertical planes, to a balanced dipole of twice the monopole's length placed in free space.

The monopole itself can take several forms. The IBA has used insulated base-fed guyed masts at the majority of ILR stations because this type of radiator is both economic to install and requires little maintenance. As a single-channel radiator it provides a reasonable load impedance when over about 60 electrical degrees high (where 1 wavelength = 360 degrees). At lower heights, sideband cutting can become noticeable and seasonal ground conductivity changes can lead to impedance instability.

Self-supporting tower radiators have not been used by the IBA since their capital cost is generally considerably higher than that of a mast of comparable height. The fact that a tower occupies a much smaller site area than a mast and its stayblocks is of little advantage since both require the installation of a large copper-wire earthmat, typically 100 metres in

diameter, if the aerial is to be both efficient and stable. Towers may well be preferable to masts when a dual or multi-channel aerial is required since their larger cross-section results in their load impedance being less frequency sensitive.

Another variation on the monopole theme that has been used by the IBA is the top-loaded base-fed mast. The connection of conductors to the top of the mast column – usually provided by uninsulated stay sections – is a useful means of making the mast appear electrically longer than its physical length. This type of radiator has been used in cases where an existing, unloaded, structure has been modified to enable it to be more suited to a lower transmission frequency.

A few shunt-fed aerials have been used by the IBA, on occasions where their use has been financially attractive. In this type of aerial the top of an earthed mast is fed by a twin or triple-wire fan tensioned down towards an area a few metres away from the mastbase. This arrangement is analogous to feeding one half of a folded dipole. However, in the MF case, the top of the mast is usually top-loaded to compensate for an electrically short structure. The physical appearance of a mast radiator surmounted by a radial array of top-loading wires has led to this type of aerial being generally described as an 'umbrella'. A shunt-fed umbrella can be adjusted to work well at a single channel but its impedance characteristic shows that it is a relatively narrow-band device, not well suited to multi-channel operation.

Tee Aerials

Typically, a 'Tee' aerial consists of a pair of vertical copper wires, spaced a few feet apart, feeding the centre of a similar pair of conductors suspended horizontally between, but insulated from, a pair of supporting masts. The masts themselves are normally open-circuited at their bases, being supported on base-insulators. This minimises the flow of RF current in the mast columns themselves – assuming they are electrically short, which is normally the case. The Tee aerial produces a nominally omnidirectional horizontal radiation pattern, and the groundwave radiation is vertically polarised.

RF currents flowing in the horizontal conductors produce horizontally polarised radiation which is mainly lost skywards, however, the resultant reduction in groundwave signal strength is of minor importance.

This type of radiator was used effectively at the temporary MF ILR station at Lots Road in central London in 1973, prior to the completion of the permanent station at Saffron Green, near Barnet. At Lots Road the ends of the Tee were supported from existing brick buildings – one either side of a small backwater of the River Thames – the water of which provided an excellent earthmat for the aerial system.

As an MF radiator, the Tee aerial has the major advantage of requiring a much-reduced height compared with that needed by an unloaded mast or tower radiator. The electrical top-loading produced by the horizontal conductors at the top of the Tee dramatically shifts the current distribution on the aerial. The result is that the base input impedance characteristic of the radiator is shifted downwards in frequency compared with that of a mast of the same overall height. Figure 1 demonstrates this point. It can be seen that although the resistance and reactance curves are similar in shape, the zero-reactance frequency for a 45m mast is about 1370kHz compared with 820kHz for a typical 45m Tee aerial as installed by the BBC at Bournemouth and Torbay and shared by the IBA. This particular 45m Tee aerial (substantially top-loaded with an 85m long top section) can be compared with a 75m unloaded mast as far as the zero reactance crossing point is concerned.

Bandwidth of MF Radiators

As is the case at other frequency bands, medium-wave aerial elements can be broadbanded by 'fattening' the conductors. Although the base input resistance of a vertical radiator is largely independent of its effective radius for heights up to about 140 electrical degrees, both the magnitude of the base reactance and the reactance/frequency slope are reduced by increasing the effective radius of the aerial.

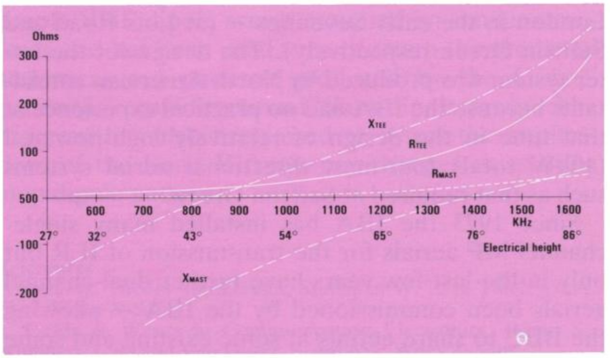

Fig.1. Base impedance characteristics of a short unloaded radiator and a 'Tee' aerial of the same overall height.

For single-channel working the choice of radiator is not of great importance since any unloaded radiator between 60 and 140 degrees high – or a considerably shorter Tee aerial – can be matched to 50ohms and still provide an acceptable load impedance over the bandwidth occupied by a modulated MF transmission. However, for multi-channel operation, often over more than an octave, it is of prime importance that the impedance of the aerial changes as slowly as possible with frequency if reasonable bandwidth figures are to be achieved, ie the VSWR is kept lower than 1.6:1 across each channel.

Unloaded mast or tower radiators can usually be made to perform satisfactorily at two channels by careful choice of height but the higher radiation resistance produced by a Tee aerial at low MF frequencies often provides a better starting point for a multi-channel aerial used over a wide frequency band. Thus the BBC have used many Tee aerials at their MF stations where they are often designed from the outset to operate on up to five channels. In contrast, the single-channel requirement of IBA stations has meant that simple mast radiators, with their economic and maintenance advantages, have normally been used.

At those IBA stations where the BBC have requested shared aerial facilities – usually after the IBA station has already gone into service – it has been possible to adapt the existing aerial system to operate satisfactorily over the two particular channels.

Performance Optimisation of Dual-channel MF Aerials

MF aerial systems designed to operate on more than one channel have an inherently poorer VSWR response over each channel than that of a single-channel aerial, due largely to the increased number of frequency sensitive networks in each transmission chain. Nonetheless, careful circuit design can minimise the degradation.

Parallel LC rejector circuits normally are used to isolate the two transmitters. They introduce either a net positive or negative reactance at the pass frequency depending on the relative disposition of the pass and reject frequencies. In all cases, however, the reactance slope (ie the rate of change of reactance with increasing frequency) introduced by the rejector across the pass channel is positive.

In addition, the impedance of a conventional base-fed mast or tower radiator has resistance and reactance slopes which are positive up to electrical heights approaching half-wave resonance, in practice at about 160 electrical degrees. All the IBA's MF radiators, with the exception of the single-channel aerial serving Edinburgh, are less than 135 degrees high consequently all the IBA's dual-channel aerials fall within this category.

The net effect is that th: reactance slopes of both the aerial and rejector al 'ays add across the pass channel. Although the impedance matching network is designed to match precisely the input impedance at the carrier frequency, the series matching components do nothing to compensate this slope, indeed they add to it.

To summarise, high resistance and reactance slopes across the pass channel degrade the overall VSWR performance, the factors which increase these slopes being as follows:

1) Aerials with steep impedance characteristics – caused by electrically thin elements and/or shunt-feeding.
2) Rejectors with large L to C ratios.
3) Operation at comparatively close frequencies.

The other major factor determining the final bandwidth performance of a dual-channel aerial is its actual radiation resistance. If this is low due to the radiator being electrically short, the matching network required to transform the impedance up to 50 ohms has the effect of magnifying the impedance spread across the channel, roughly in proportion to the transformation ratio. Short radiators are economically desirable, however, and are more likely to gain the approval of Planning Authorities than tall structures.

Case Study of the Dual-frequency Aerial Used at East Kent

A relatively low frequency of 603kHz was chosen for the omnidirectional aerial required to cover the whole of the East Kent ILR service area from a site just outside Canterbury. A considerably lower-than-ideal mast height had to be used in order to obtain planning permission for the station which was also to be shared by the BBC on 774kHz.

The structure chosen was a base-fed mast 81 metres high, triangular in cross-section and with a slim outline due to the 0.39m facewidth. Electrical top-loading was not used. This height corresponds to 58 electrical degrees at 603kHz and 75 degrees at 774kHz.

When designing an aerial system of this kind it is desirable that reasonably accurate base-impedance data are available for the proposed radiator other-

wise predicted and actual aerial performances can be substantially different. An identical structure had previously been used at ILR Peterborough consequently this information was readily available.

Predicted mast operating impedances and slopes were as follows:

603kHz: $Z = 15 - j78$ ohms:
$dZ/dF = 0.12 + j0.62$ ohm/kHz
774kHz: $Z = 25 + j30$ ohms:
$dZ/dF = 0.16 + j0.70$ ohm/kHz

Using these predicted impedances as the basis for the aerial design, it became apparent that the low operating resistance at 603kHz would give rise to sideband cutting. Without any form of prematching and using rejectors with standard L/C ratios of about 10,000 Henries/Farad and Qs of about 300 to 500, a VSWR of not better than 2.7:1 could be expected at the lower channel's sideband frequencies. The performance at the higher channel would be better since the mast impedance is more favourable.

At this stage in the design electrical top-loading was considered but discounted since it was felt that, if a system with an acceptable performance could be derived by using conventional circuitry coupled to a standard radiator whose impedance parameters were already known, this would be preferable to the alternative of using a top-loaded structure. Top-loading adds considerably to the complexity of the system, there are uncertainties in its predicted performance, and the installation and maintenance costs are greater.

Premathing

As previously stated, a much improved performance at the lower channel can generally be achieved if the mast operating resistance can be transformed upwards before being presented to the output of the rejector.

In the particular case of the East Kent system, (see fig.2) this was achieved by placing an inductor in series with the aerial to increase the net positive reactance at the higher channel whilst maintaining a small negative one at the lower channel. The value of this inductance was chosen so as to keep the net operating reactance-to-resistance ratio as small as possible, ideally less than 4, at both channels. A parallel LC combination was then placed in shunt so as to transform the resistance at the two carrier frequencies to higher values, particularly at the lower channel.

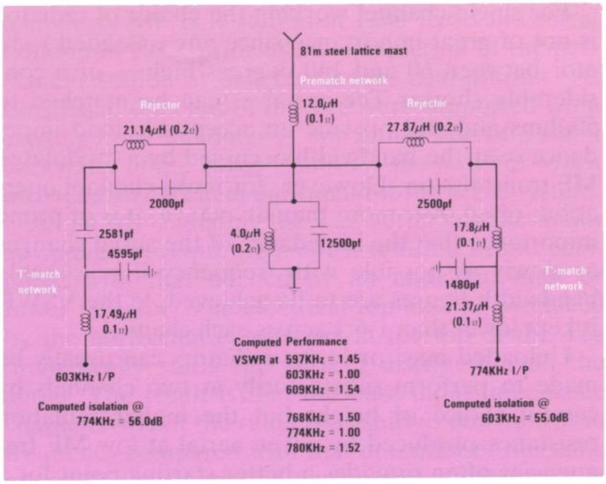

Fig.2. Combining and matching networks for the Canterbury MF aerial.

The LC component values of this parallel combination were chosen such that the network provided a net shunt positive reactance at the lower frequency and a negative reactance at the upper one.

The following formulae give exact solutions for the L and C values required to effectively resonate the reactive components of the two adjusted impedances at the high and low frequencies.

$$L = \frac{X_L}{\omega_L (1 + \omega_L X_L C)}$$

$$C = \frac{\omega_L X_H - \omega_H X_L}{X_L X_H [\omega_L^2 - \omega_H^2]}$$

where X_L and X_H are the required reactances to be provided by the LC network at the lower and upper frequencies respectively. ω_L and ω_H relate to the lower and higher frequencies.

The calculated component values tend to be somewhat impractical since they can involve large values of capacitance and low inductance which, in turn, lead to high currents being driven through them to earth producing appreciable heat losses. Compromise solutions are normally found where bandwidth performance is much improved whilst limiting the associated circulating currents in the LC components. At this stage, much use is made of computer programs which analyse the performance of the possible circuits in the presence of stray capacitances which inevitably occur in practice.

The shunt LC combination is a low-Q circuit the resonant frequency of which lies between the two

operating channels. Components situated between the rejectors and the radiator constitute the prematch section.

In the case of the East Kent aerial, the use of prematching enabled the lower channel sideband VSWRs to be reduced from 2.7 to 1.6 whilst still achieving the same VSWR figure of 1.6 at the limits of the upper channel; the circuit used is shown in fig.2.

Prematching usually reduces the isolation achieved between the two inputs but this is not normally of significance provided that there is still sufficient isolation to prevent the formation of intermodulation products within the two transmitters. High circulating currents can be a problem in prematch sections since the reactances of their individual components tend to be low. However by the use of computer programs it is possible to derive the necessary peak voltage and RSS current ratings of components and verify the network losses and isolation between transmitters.

Actual measurements made on the completed aerial system at Canterbury were in close agreement with its predicted performance.

Filter Requirements in MF Aerial Systems

In any aerial system which is designed to radiate more than one channel, it is necessary to insert filters to prevent cross-modulation between the transmitters and to minimise the levels of any intermodulation products. These filters also permit independent control of matching networks. In the case of directional MF aerial arrays – where the radiating elements are driven with unequal currents at different phases in order to achieve the desired radiation pattern – filters also permit separate power division and enable phasing networks to be set up independently for each channel.

Bandpass or indeed high or low pass filters are generally not suitable for MF networks since they would need too many sections to achieve the necessary rejection at the unwanted frequencies. The required isolation between transmitters is usually provided by the insertion of a series of precisely-tuned notch rejectors, with one rejector for each unwanted frequency, in each transmission chain. In the case of a dual-channel aerial, this becomes a pair of filters, located at the base of each radiating element – one in each transmission chain.

Filter Types

Several types of filter can be used in MF networks – four of which are shown in fig.3. It is interesting to

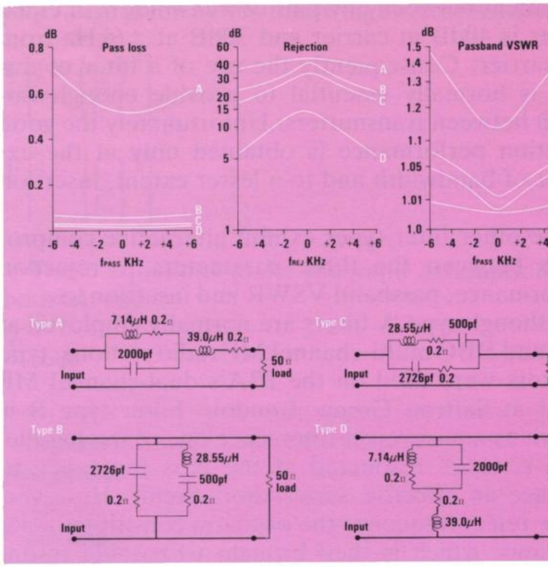

Fig.3. Filter configurations used at dual channel MF stations.

compare the relative performances of these various types so that the most appropriate one can be chosen for any particular application. In an attempt to gain a true performance comparison between the filter types, resistive losses have been kept constant with values typical of those measured in practice.

Each filter type has both advantages and disadvantages over the others – as can be seen by studying the curves shown in fig.3. (The filters are shown working into a 50 ohm resistive load, for comparison purposes.)

The most commonly used filter in multi-channel MF aerials is type A. This comprises a simple parallel LC rejector in series with either an inductor or a capacitor depending on whether the pass frequency is higher or lower than the reject frequency (in the case shown the pass frequency is higher). The reactance of the compensating series inductor cancels out the net capacitive reactance produced by the parallel combination at the pass carrier frequency but, unfortunately, increases the reactance slope across the pass channel.

In practice, the compensating component is not actually seen in the form of a separate inductor or capacitor but is merged into the output arm of the Tee-match network, the 'filter' then becoming a simple parallel LC rejector.

The primary advantage of this type of filter is the high level of rejection it provides across the com-

plete stopband. The normal minimum isolation requirement between any pair of transmitters of equal power is 40dB at carrier and 30dB at ±6kHz from the carrier. Consequently the use of a filter of this type is normally essential to provide enough isolation between transmitters. Unfortunately the good rejection performance is obtained only at the expense of bandwidth and to a lesser extent, insertion loss.

The other filter types exhibit alternative compromises between the three parameters – rejection performance, passband VSWR and insertion loss.

Although type A filters are normally employed at IBA and BBC multi-channel MF radio stations, type B filters were used on the IBA's dual-channel MF aerial at Saffron Green, London. Filter type B is known as a pass/reject filter since one of the parallel arms is series resonated at the pass frequency to produce an effective series short-circuit. However, at the reject frequency the pass arm constitutes a net reactance which is then brought to parallel resonance by the other arm consisting of an inductor or capacitor depending on whether the reject frequency is below or above the pass frequency. This pass/reject filter gives excellent passband performance – particularly important in multi-mast arrays if the radiation pattern is to remain constant over the audio bandwidth of the transmission – with somewhat modest rejection between transmitters. However, this was considered to be acceptable at Saffron Green since the valve transmitters were reasonably tolerant of interfering carriers.

Pass/reject filters provide less rejection than a type A rejector for a given value of loss resistance and their rejection response is much narrower, falling off much more rapidly in the sidebands, since their dynamic resistance is rather low despite having a high Q. The rejection they provide is normally inadequate when solid-state transmitters are in use and therefore they are not widely used in the UK.

Type D filters were employed at the Peterborough ILR station to provide additional isolation between the 1332kHz ILR and 1449kHz BBC transmitters. The 40dB isolation provided by a pair of standard rejectors (type A filters) proved insufficient to prevent the generation of two 3rd-order intermodulation products by the solid-state transmitters when operating at this close-frequency spacing. (This relatively low primary-rejection figure was due largely to the high operating impedances involved at the two carrier frequencies.) The insertion of a type D filter across the input to each impedance matching circuit provided a further 7 to 10dB of rejection without significantly affecting the input bandwidth of each channel.

Factors Affecting Rejector Performance

The absolute performance of a simple parallel rejector (filter type A) is determined by its L/C ratio and by the resistive losses of the components and their interconnections within the resonant circuit. The 'Q' of the resonant circuit gives a good indication of the rejection performance to be expected of a filter. However it does not tell the whole story. For example, two rejectors can have the same Q whilst having different L/C ratios by having different loss resistances but their rejection characteristics will not be the same. In all cases the circuit with the higher L/C ratio will provide increased rejection across the whole stopband. As far as rejector performance is concerned, the following rules apply:

1) Rejection at sidebands is far more dependent on the L/C ratio of the circuit than on its Q, rejection increasing with increasing L/C ratio.
2) The rejection achieved at the precise reject frequency increases with both increasing L/C ratio and decreasing circuit losses.

These properties are clearly demonstrated in fig.4 for three examples where Q and L/C are varied.

Limitations of Parallel Rejectors

Rejectors produce heat owing to the passage of currents at both pass and reject frequencies through the resistive losses contained within the resonant circuit. In addition, power can be lost in eddy currents flowing in the walls of the filter enclosure, consequently it is normal practice to make the enclosure of aluminium and to make it as large as practicable so as to minimise such losses.

Reject-frequency currents can be reduced in the rejector by increasing the L/C ratio, albeit at the expense of increasing the pass-frequency VSWR, and by minimising the operating voltage of the aerial at the reject frequency, since the reject frequency currents are effectively voltage driven.

In a type A filter, when the pass and reject frequencies are widely spaced, the pass frequency current driven into the operating impedance of the aerial is virtually all carried by only one of the components of the parallel LC filter. However, as the frequency separation decreases the currents through the two components tend to become equal in amplitude (but always opposite in phase). Only the net current flowing through the parallel combination

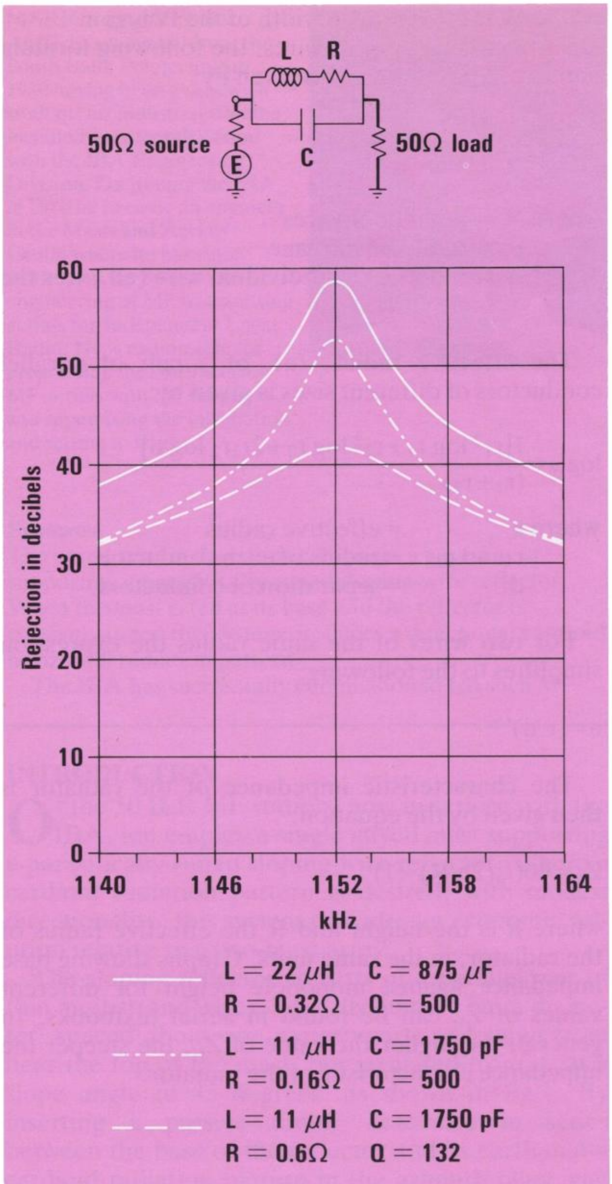

Fig.4. Variation in rejection performance of a parallel rejector for differing L/C and Q values.

feeds power into the load impedance of the aerial, consequently substantial pass-frequency circulating currents result when close frequency spacings are used.

It is interesting to note that this current amplification is independent of Q, L or the L/C ratio. It is dependent only on the ratio of the reject and pass frequencies. Currents flowing in the inductor (I_L) and capacitor (I_c) compared with the net pass current I_p are given by the following two expressions:

$$I/\left[1-\left|\frac{f_p}{f_r}\right|^2\right]$$

$$I/\left[1-\left|\frac{f_r}{f_p}\right|^2\right]$$

These equations make the valid assumption that both ωL and

$$\frac{1}{\omega_c} \gg R.$$

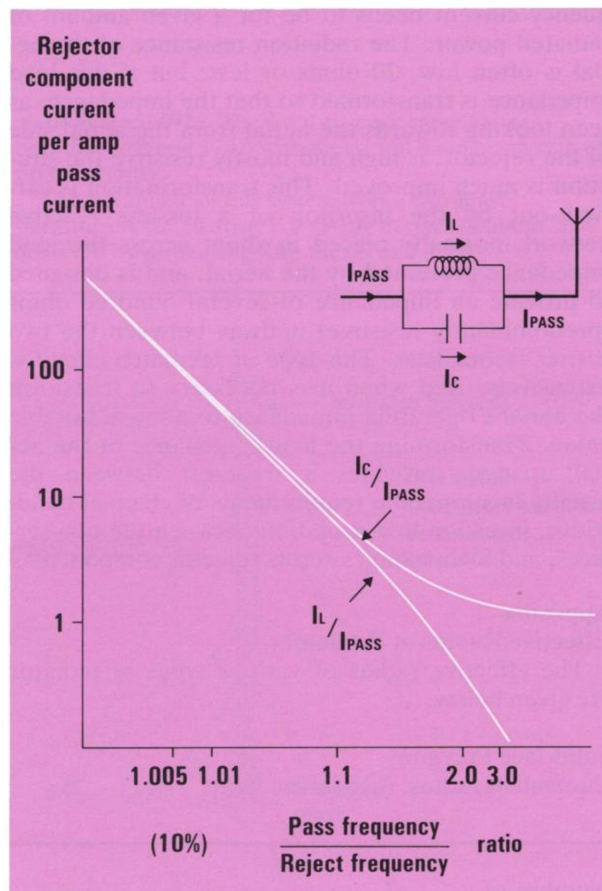

Fig.5. Variation of pass-frequency circulating currents with frequency separation for a type A filter.

Figure 5 shows a plot of the two characteristics. It can be seen that pass-frequency circulating currents increase dramatically with decreasing frequency sep-

aration. In many aerial systems, these pass-frequency currents can often be higher than the reject-frequency resonant currents. Both absorb appreciable amounts of transmitter power, dissipated as heat, and produce high peak voltages across the circuit components.

Close Frequency Operation

Frequency separations of approximately 10% or more present no particular problems, however special steps need to be taken with closer frequency separations if significant insertion losses and high component voltages are to be avoided.

It follows that the higher the operating resistance of the aerial, the lower the rejector output pass-frequency current needs to be for a given amount of radiated power. The radiation resistance of the aerial is often low, 20 ohms or less, but if this load impedance is transformed so that the impedance, as seen looking towards the aerial from the aerial side of the rejector, is high and mostly resistive the situation is much improved. This transformation is carried out by the insertion of a lossless reactive network normally placed in shunt across the load impedance presented by the aerial, and is designed to provide an impedance of several hundred ohms (predominantly resistive) midway between the two carrier frequencies. This type of prematch circuit is extensively used when it is necessary to transform the aerial's operating impedance to a more suitable value. Transforming the load impedance of the aerial upwards provides a trade-off between the usually incompatible requirements of channel bandwidth, insertion loss, isolation between the two services, and high voltages across rejector components.

Appendix

Effective Radius of Radiators

The effective radius of various types of radiator are given below.

Solid-face Polygons
Equivalent radius, (a) is given by:

Number of Sides	2 (flat strip)	3 (triangle)	4 (square)	5	6
Equivalent Radius	0.25W	0.42W	0.59W	0.76W	0.92W

where W is the one-face width of the Polygon.

For a circular cage of wires, the following formula applies. Equivalent radius given by:

$$a=R\left[\frac{N.r}{R}\right]^{1/N}$$

where N = number of wires
R = radius of the cage
r = radius of an individual wire (all wires the same size).

The effective radius, (a), of a pair of parallel conductors of different sizes is given by:

$$\log a=\frac{1[r_1^2 \log r_1+r_2^2 \log r_2+2r_1r_2 \log d]}{(r_1+r_2)}$$

where a =effective radius
r_1 and r_2 =radius of each conductor
d =separation of conductors.

For two wires of the same radius the expression simplifies to the following:

$$a=(r.d)^{1/2}$$

The characteristic impedance of the radiator is then given by the equation:

$$Z_o=60(l_n(2h/R)-1)$$

where h is the height and R the effective radius of the radiator, in the same units. Graphs showing base impedance against monopole height for different values of Z_o can be found in aerial textbooks. In general, the higher the value of Z_o, the steeper the impedance characteristic of the radiator.

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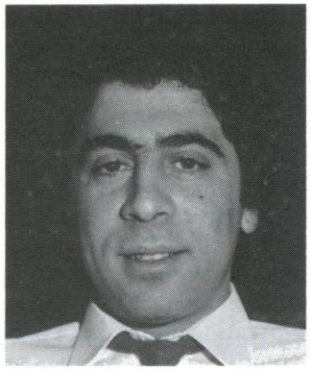

Directional MF Aerials With Sloping Wire Reflectors

by M. Zapitis

Synopsis

The aerial system described comprises a guyed mast supporting a parasitically-tuned sloping-wire reflector. When the mast is fed at its base and the reflector is passively tuned this system produces a directional cardioid horizontal radiation pattern.

The IBA has successfully commissioned ten such MF stations since 1975. The front-to-back ratios range between 5dB and 17dB with measured main beam gain of up to +2.5dB.

The advantage of this system over a two-mast parasitically tuned array is a saving in costs of about 20% achieved by dispensing with the second mast.

INTRODUCTION

Of the 50 ILR MF stations now in service with the IBA, ten employ a single guyed mast supporting a parasitically-tuned sloping-wire reflector. Where a cardioid radiation pattern is desired, with modest directionality, this system provides an economic solution relative to a two-mast array.

The sloping reflector consists of a parallel pair of thin multistrand wires, typically 3mm to 6mm diameter, spaced about one metre apart, and slung from near the top of the mast, but insulated from it, at a slope angle of 45 degrees, as shown in fig.1. By inserting a passive tuning reactance in series between the base of the reflector and its earth mat a cardioid radiation pattern in the azimuth plane can be achieved with a maximum front-to-back ratio typically of around 13dB to 17dB. Measured forward gains of up to +2.5dB over the theoretical short monopole normally used as a reference for MF aerials have been achieved. To minimise ground-system losses both the mast and the sloping wire have their own radial-wire earth mats. These usually consist of 72 or 90 buried copper wires per radiator extending radially outward from the base of the radiator to a minimum distance of λ/6. The electrical height of both the mast and the reflector are important in

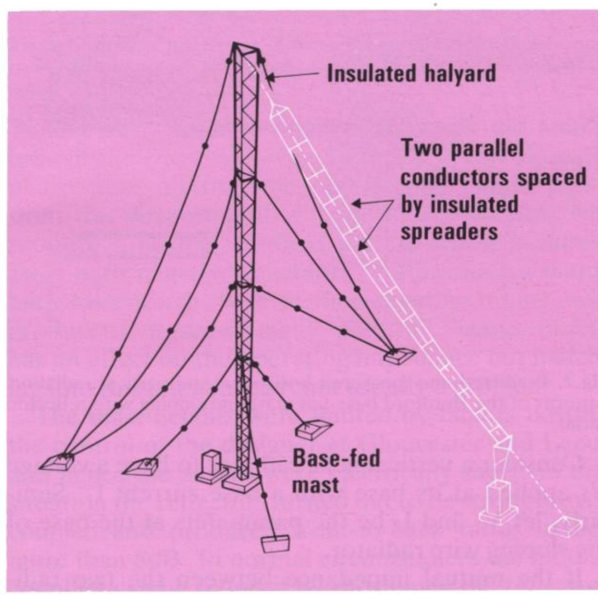

Fig.1. A directional MF aerial with base fed mast and sloping twin wire reflector.

determining the amount of directionality achieved in the horizontal radiation pattern. The most geometrically 'compressed' aerial, at Gloucester (60°

mast, 72° reflector, where one wavelength = 360°), exhibited a mere 8dB front-to-back ratio with a forward gain of −1.0dB, whereas at Coventry the use of a taller mast (93° mast, 84° reflector) led to a maximum front-to-back ratio of 24dB with a forward gain of +2.4dB being measured.

Theoretical calculations show that the sloping-wire radiator can not be operated as a director, with more than about 2.0dB front-to-back ratio, whatever value of tuning reactance is inserted in series with it.

Theoretical Studies

The first aerial to be built on this principle for ILR was at the Bradford MF station in 1975. The system was developed from an empirical laboratory model scaled to 1/60 size. The measured radiation pattern of the finally-installed full-scale aerial is shown in fig.2. Measurements made at Bradford were subsequently used to set up a computer model to provide insight into how the reflector operated and to predict its performance at future stations.

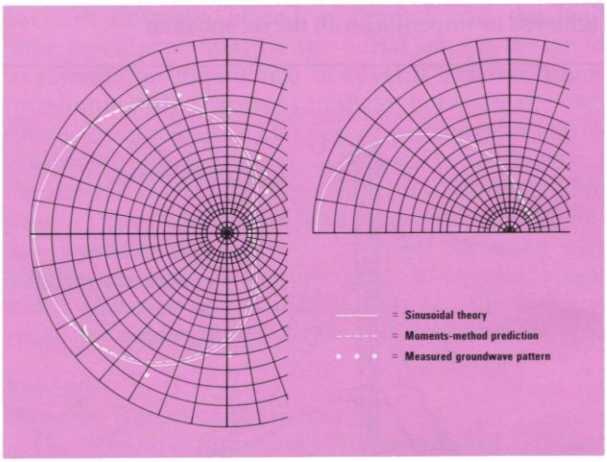

Fig.2. Predicted and measured horizontal and vertical radiation patterns of the Bradford base-fed mast and sloping wire reflector aerial.

Consider a vertical mast radiator to have a voltage V_1 applied at its base with a base current I_1. Similarly, let V_2 and I_2 be the parameters at the base of the sloping wire radiator.

If the mutual impedance between the two radiators is Zm then:

$$V_1 = I_1.Z_{11} + I_2.Zm \quad (1)$$

$$V_2 = I_2.Z_{22} + I_1.Zm \quad (2)$$

Z_{11} and Z_{22} are the base self impedances of the mast and sloping reflector respectively. If the reflector is now earthed at its base, then V_2 in equation (2) becomes zero, i.e.;

$$I_2 = -I_1.Zm/Z_{22} \quad (3)$$

If the parasite's base is earthed through a series impedance, equation (3) still applies provided Z_{22} is the sum of the reflector's self impedance plus the tuning impedance. The vector equation (3) defines the amplitude and phase of the current I_2 at the base of the tuned parasite in terms of the fed radiator's base current I_1, the negative sign implying 180° phase change. Substituting from (3) into (1) the operating input impedance of the fed element becomes:

$$Z_1 = Z_{11} - (Zm)^2/Z_{22} \quad (4)$$

and similarly for Z_2 if the mast were to be made the parasite. The base mutual impedance was derived using the induced EMF method, by integrating the near electric field component along the parasitic element produced by the radiator and its image in the ground plane, assuming a sinusoidal current distribution on the mast and the parasitic reflector.

Moments Method Techniques

All the early designs were based on the sinusoidal current distribution theory. For a finite diameter conductor the sinusoidal current distribution is representative but not exact. Recently a computer model has been set up, using a technique based on the numerical Moments Method, to investigate the design of mast and sloping wire antenna systems with variable geometries.

The aerial system including its ground image is sub-divided into N segments, where N is typically between 20 and 200 depending on the complexity of the model. The matrix equation $V=I \times Z$ is generated by piecewise-sinusoidal current expansion and the impedance matrix (order $N \times N$) is defined and calculated by either numerical integration using Simpson's Rule or by closed-form impedance expressions in terms of exponential integrals. The impedance matrix is then inverted and solved for the current distribution knowing the voltage at the input terminals. Having determined the current distribution, the input impedance and radiation patterns can then be obtained.

While the current magnitude bears a general resemblance to the sinusoidal distribution there are noticeable differences between the two, especially near the base of the radiators. Such differences may be significant in calculating the near fields and the base operating impedance. Studies carried out on the ten aerials in service indicate that the sinusoidal

current distribution assumption is accurate for conductor lengths of up to $\lambda/4$, but differences become significant when the lengths depart from this. The horizontal radiation pattern shown in fig.2 demonstrates the difference in the two methods and compares them to the measured patterns. Moments method predictions are in close agreement with measurements taken on site.

Aerial Performance Analysis

Although a careful choice of geometrical arrangement can result in front-to-back ratios as high as 25dB it should be noted that the groundwave horizontal radiation pattern follows a well defined shape, ie. the null to the rear does not form a narrow notch but is maintained over a broad azimuth arc to create what may be described as an "apple-shaped" pattern. The suitability of a sloping-wire reflector system at any station would depend on the acceptability of this pattern shape.

It has been discovered that in both the Moments method and sinusoidal theory an empirical earth loss resistance had to be introduced at the bases of the radiators in order to achieve the front-to-back ratio measured on site. The lossless case predicts that the front-to-back ratio will not exceed 10dB for any aerial arrangement.

For a given pattern shape and front-to-back ratio, there are several parameters within the control of the designer. These include the electrical height of the mast and length of the reflector, their electrical spacing and the reflector's base tuning reactance. Figure 3 shows the effect of the reflector tuning reactance versus MF gain for various reflector lengths, including the length actually used at the Coventry ILR station. By inspecting equation (3) it is seen that the reflector current is proportional to Z_M/Z_{22}. A careful choice of aerial geometry and reflector tuning reactance can control the amplitude and phase of the reflector current to produce a range of null depths. At Coventry, for instance, the maximum front-to-back ratio occurs when the current in the reflector is 30% greater in amplitude than the mast current. This is because the mast generates 30% more radiated field than the reflector for a given current at its base, and the favourable phase-centre spacing between reflector and mast results in almost complete field cancellation to the rear.

For a given sloping-reflector length it can be seen from fig.3 that the null-depth is adjustable (at Coventry this was 9dB in operational service) by adding passive tuning reactance at the base of the reflector.

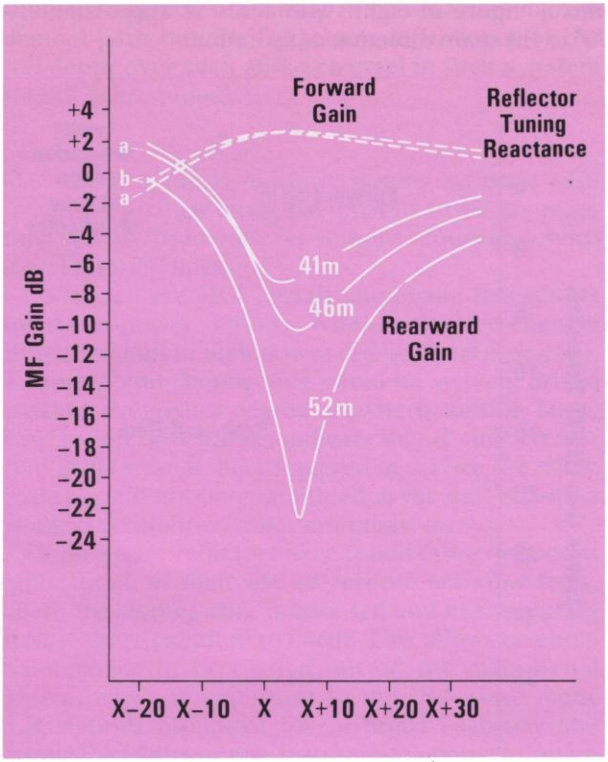

Fig.3. Computer prediction showing the variation of MF gain with reflector tuning for various reflector lengths for the Coventry MF aerial.

Tuning away from resonance decreases the amplitude of the current flowing in the wire and alters its phase, thus altering the null depth. Tuning away from the deepest null with positive reactance will produce a normal cardioid pattern whereas tuning away with negative reactance will produce a major back lobe which, with further negative tuning, will produce a 'figure-of-eight' pattern. Passive tuning has an effect on the operating impedance but matching into this impedance is not usually a problem.

The mast heights were limited by factors beyond the control of the designer at Gloucester and Leeds and hence the aerials were electrically short. In this situation the reflectors turned out to be strongly over coupled and produced front-to-back ratios of not more than 8dB. In normal circumstances the mutual coupling can be reduced by shortening the reflector, but when the reflector is already as short as 67° a reduction in its length causes the self-resistance to drop at the same rate as the mutual impedance, thus the overcoupling remains. When this was tried at Leeds (mast 60°, reflector 67°) the pattern developed

into a 'figure-of-eight' with nulls at approximately 90° to the main direction of radiation.

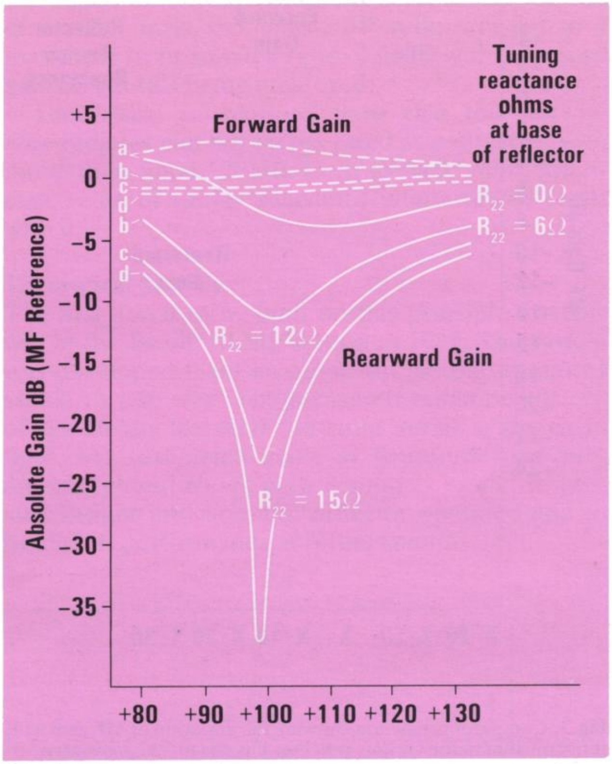

Fig.4. Computer predictions showing the variations in forward and rearward gain for Gloucester MF aerial with differing values of earth mat loss resistance.

Another way that the current in the reflector can be attenuated without changing its phase is by inserting additional loss resistance at its base. Figure 4 shows the theoretical effect of adding varying amounts of loss resistance to the base of the reflector at Gloucester. As seen from the diagram very deep nulls could be obtained at the expense of forward gain since some of the transmitter power is dissipated as heat in the resistor. At three stations with short masts it was necessary to resort to a 6-ohm resistor in series with the variable tuning reactance. Although this may seem an undesirable feature, it did not create any problem for the IBA because the transmitter powers were very low and had been specified originally to deal with low efficiency antennas. For high power, high efficiency requirements, resistive damping would be unacceptable. In more recent installations, reflectors with adjustable lengths have been used in an attempt to overcome the problem.

All ten stations using the sloping reflector configuration have nulls ranging between 5dB and 17dB in depth. These nulls are not intended to provide co-channel protection, but are intended to adjust the size of the service area to conform with the desired editorial boundaries, consequently their stability has not been monitored. There is no reason to suspect any instability, however, in this type of aerial system.

The feed-point VSWR of the aerials at ±6kHz from carrier varies from 1.05 for stations with front-to-back ratios of 5dB, such as at Bury St. Edmunds, to 1.30 for systems with front-to-back ratios of 17dB, such as at Bradford and Preston.

Sloping-Wire Radiators

Sloping-wire conductors can also be used as radiators. A mast with a sloping-wire reflector has a reserve transmission capability inherently built into it . If, for example, the mast or its networks need to be maintained, the sloping radiator can be used as

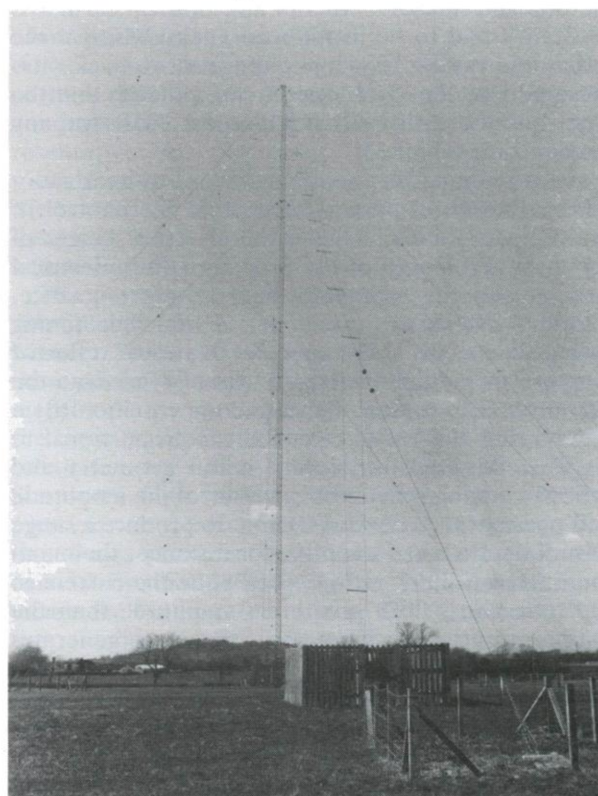

Fig.5. The base fed mast radiator and twin sloping wire reflector in use at the Gloucester MF station.

the fed element whilst the mast is short-circuited to earth. Under these conditions the directionality of the system is changed and it becomes nominally omnidirectional if the electrical height of the mast is substantially less than a quarter-wavelength (ie. the mast is non-resonant). For mast heights greater than a quarter-wavelength some directionality remains.

Indeed, sloping-wire radiators are used as reserve transmitting systems on all IBA MF radio stations. At London and Birmingham, for example, there are dedicated dual-channel reserve sloping wire aerials used to radiate the services when maintenance is necessary on the main array. Under these circumstances the four masts are short-circuited to earth to make them safe to climb and the sloping radiator is fed at its base, the result being a directional cardioid pattern with a forward gain of 1 to 2dB and maximum front-to-back ratio of 8 to 10dB.

Dual Channel Operation

Recently the Bedford aerial was converted to dual channel operation. The existing channel had an operational front-to-back ratio of 14dB and a forward gain of +2.4dB at 792kHz. The channel to be added was 1161kHz. Computer studies had shown that the sloping reflector would be considerably off-tune at 1161kHz and carry only a small base current relative to the mast current, producing an almost omnidirectional horizontal radiation pattern with 2dB front-to-back ratio which in this instance was acceptable for the 1161kHz channel. The measured pattern was in very close agreement.

In cases where directionality is required on both channels, the reflector could be independently tuned at both frequencies by fitting rejection filters in each channel path but the frequency-dependence of the null depth over each audio channel in such a system has not been studied.

Conclusions

The use of a tuned parasitic sloping reflector supported by a single mast has proved to be an economic system for low-power stations requiring service area 'editorial' nulls.

So far, all ten IBA aerials employing this system are at low-power stations. Where a deep co-channel protection null in high power applications is desired, both mast and sloping wire could be actively driven to achieve a greater degree of pattern control. However, theoretical studies indicate that if nulls greater than 25dB were to be attempted in service the audio quality could not be guaranteed in the null region, a problem common to other multimast arrays.

Sloping-wire radiators have found other important applications in their use as reserve aerial systems, where the sloping-wire is base fed and its supporting mast is short-circuited to earth. This allows essential maintenance to be carried out on the main aerial system, which is important to the IBA since many ILR stations broadcast for 24 hours each day and Safety Regulations put restrictions upon the maintenance of live MF masts.

From the planning point of view a sloping-wire system, fig.5, is more acceptable than a two mast array because it has the appearance, from a distance, of a single mast – the reflector being scarcely more obtrusive than the guy wires.

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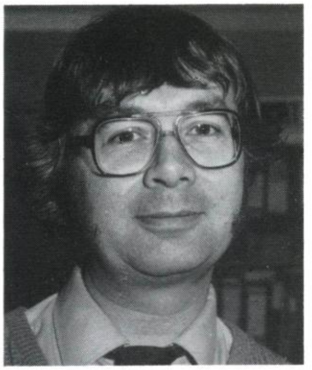


Multi Channel Impedance Matching At UHF And VHF

C. K. Jacob

Synopsis

The requirement of the UK Television Act, 1963, to co-site IBA and BBC transmitters for the UHF national television network has led to the use,at most stations, of a single transmitting aerial capable of radiating all four services simultaneously. Novel transmission line matching techniques have been evolved to overcome the problems of satisfying the stringent VSWR limitations necessary in such systems, in order to reduce the radiation of delayed images, and some of these techniques, which also have a general application, are developed. This is followed by an outline review of broad-banding methods and the article concludes with a discussion of some of the practicalities associated with impedance measurement and correction at RF.

INTRODUCTION

This article is primarily concerned with matching procedures applicable to passive high-power RF systems which employ coaxial distribution feeders. Coaxial systems lend themselves to correction methods that utilise simple shunt-connected reactances, as these may be easily realised in practice by means of metal sleeves, (or undercuts) on the inner conductor. It is for this reason that the article concentrates on techniques that make use of simple parallel-connected lumped susceptances, although all formulae quoted are equally applicable to series connected reactances provided that Voltage Reflection Co-efficient, (ρv) is substituted for Current Reflection Co-efficient (ρi).

The Appendix contains a summary of basic transmission line formulae.

Effect of Multiple Discontinuities on a Transmission Line

It is necessary to understand how several discontinuities, each modelled as a pure shunt susceptance, together affect the input reflection co-efficient of an otherwise uniform transmission line.

An accurate determination requires that the effect of individual susceptances be added to the line admittance at each appropriate position while referring from load to input. It is clear that:

(i) the way that several mismatches combine is frequency dependent

(ii) large discontinuities will interact in a non-linear manner.

A simplification is possible where load and line mismatches are relatively small, since reflection co-efficients may then be combined vectorially to give a good approximation. The basis of this approximation is the linear nature of the transformation from reflection co-efficient to impedance or admittance co-ordinates. Figure 1 shows a polar reflection co-efficient chart overlaid with lines of constant conductance and susceptance in normalised units. Note that the central region of the chart within the marked

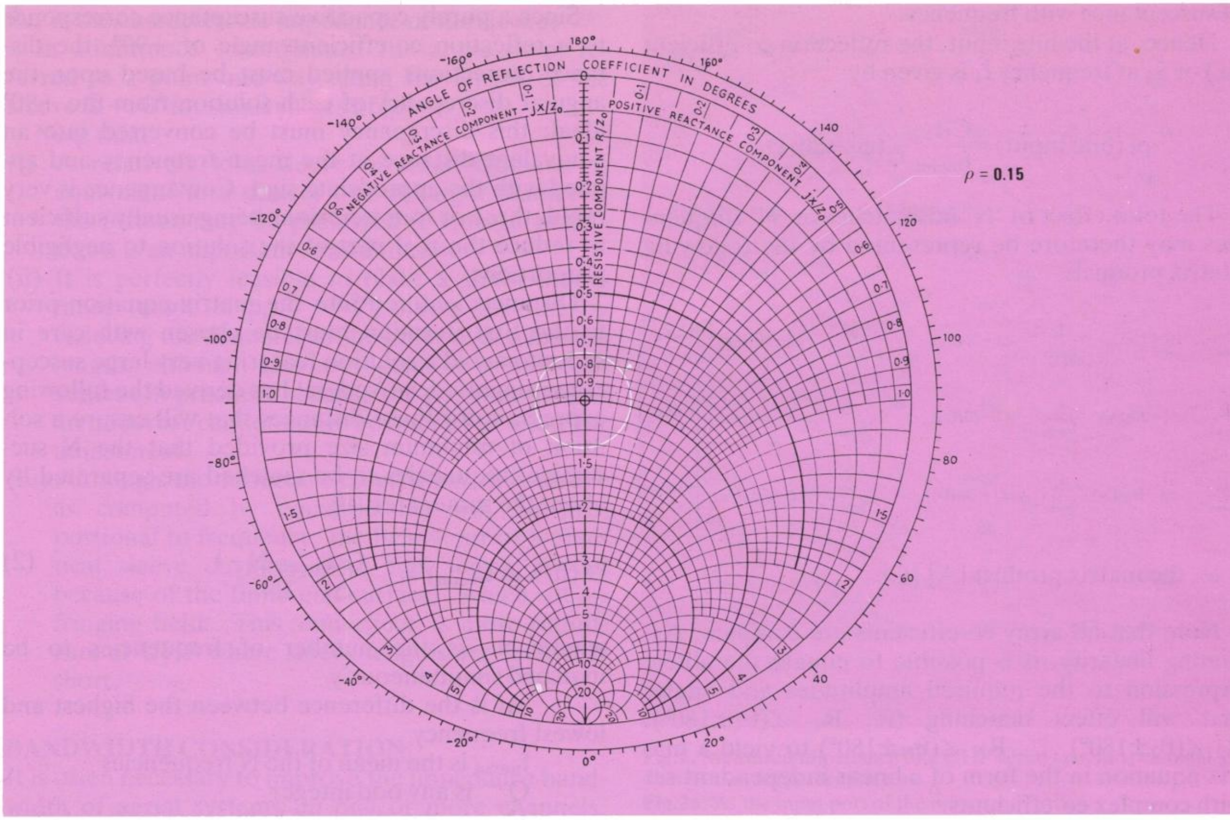


Fig.1. In the area of the Smith chart lying within the circle corresponding to $\rho = 0.15$ the grid is sufficiently linear to allow vectorial manipulation.

circle, corresponding to ρ magnitude values smaller than 0.15, comprises an approximately linear grid, thus justifying vectorial manipulation in this area.

Multiple Frequency Matching

The above-mentioned linear approximation is the basis of a line technique that will simultaneously match several specified points on the characteristic of a frequency dependent load.

Consider a lossless line terminated in a mismatched load such that the complex current reflection coefficient (ρi) at the line input is:

$R_1 \angle\theta_1$ at frequency f_1
$R_2 \angle\theta_2$ at frequency f_2
. .
. .
$R_N \angle\theta_N$ at frequency f_N

If the line is to be matched simultaneously at all 'N' spot frequencies, then 'N' correcting shunt susceptances, each of a different size, must be placed across the line at distances $D_1, D_2, \ldots D_N$ respectively from the load. Each correcting susceptance can be considered, when positioned on the line, to act as a point reactive mismatch: let the reflection co-efficient (ρ_i) produced by:

$$\left.\begin{array}{lll} \text{susceptance } S_1 & \text{be} & C_1 \angle\phi_1 \\ \text{susceptance } S_2 & \text{be} & C_2 \angle\phi_2 \\ & & \cdot\ \cdot \\ & & \cdot\ \cdot \\ \text{susceptance } S_N & \text{be} & C_N \angle\phi_N \end{array}\right\} \begin{array}{c}\text{at the mean} \\ \text{frequency} \\ f_{mean}\end{array}$$

Consider susceptance S_1: Its effect at the line input may be determined by referring the reflection co-efficient $C_1 \angle\phi_1$ through distance D_1. At frequency f_1, for example, this implies that ϕ_1 changes to $(\phi_1 - 2ß_1D_1)$, where $ß_1$ is the phase constant $(2\pi/\lambda_1)$ associated with f_1. In addition, as magnitude C_1 is specified at the mean frequency, it must be multiplied by f_1/f_{mean} to allow for the change

of susceptance with frequency.

Hence, at the line input, the reflection co-efficient (ρ_i) of S_1 at frequency f_1 is given by:

$$\rho_i \text{ (line input)} = \frac{C_1 f_1}{f_{mean}} \angle(\phi_1 - 2\beta_1 D_1)$$

The total effect of 'N' mismatches at 'N' frequencies may therefore be represented by the following Matrix product:

$$\begin{bmatrix} \frac{f_1}{f_{mean}} \cdot \angle(-2\beta_1 D_1) & \frac{f_1}{f_{mean}} \cdot \angle(-2\beta_1 D_2) & \cdots & \frac{f_1}{f_{mean}} \cdot \angle(-2\beta_1 D_N) \\ \frac{f_2}{f_{mean}} \cdot \angle(-2\beta_2 D_1) & \frac{f_2}{f_{mean}} \cdot \angle(-2\beta_1 D_2) & \cdots & \frac{f_2}{f_{mean}} \cdot \angle(-2\beta_2 D_N) \\ \vdots & \vdots & & \vdots \\ \frac{f_N}{f_{mean}} \cdot \angle(-2\beta_N D_1) & \frac{f_N}{f_{mean}} \cdot \angle(-2\beta_N D_2) & \cdots & \frac{f_N}{f_{mean}} \cdot \angle(-2\beta_N D_N) \end{bmatrix} \begin{bmatrix} C_1 \angle\phi_1 \\ C_2 \angle\phi_2 \\ \vdots \\ C_N \angle\phi_N \end{bmatrix}$$

ie. the matrix product [A].[C]

Note that all array co-efficients are complex. Assuming linearity, it is possible to equate the above expression to the required amplitudes and angles that will effect matching (ie. $R_1 \angle(\theta_1 \pm 180°)$, $R_2 \angle(\theta_2 \pm 180°) \ldots R_N \angle(\theta_N \pm 180°)$ to yield a matrix equation in the form of a linear independent set with complex co-efficients:

ie.

$$[A].[C] = \begin{bmatrix} -R_1 \angle\theta_1 \\ -R_2 \angle\theta_2 \\ \vdots \\ -R_N \angle\theta_N \end{bmatrix} \quad (1)$$

Several classical techniques exist to solve these equations for [C] but substitution methods such as Gaussian elimination, are probably the quickest and most straight forward. This method may be used without any form of pivoting since there is no possibility of the magnitude of any element in matrix [A] being zero.

After solving (1) above, the computed complex values $C_1 \angle\theta_1$, $C_2 \angle\theta_2 \ldots C_N \angle\theta_N$ will be found to have arbitary phase angles, and since any non-dissipative correcting mismatch must be purely reactive, it will not generally be possible to implement the initial solution.

A practical result, representative of pure susceptance, can be achieved by repeatedly solving equation (1) with successive adjustment of D_1, $D_2 \ldots D_N$.

Since a purely capacitive susceptance corresponds to a reflection co-efficient angle of +90°, the distance corrections applied must be based upon the angular discrepancy of each solution from the +90° ideal: this discrepancy must be converted into an equivalent distance at the mean frequency and applied with the appropriate sign. Convergence is very rapid, three or four iterations being usually sufficient to reduce the real part of the solution to negligible proportions.

Distances entered into the matrix equation prior to the first iteration must be chosen with care in order to avoid a solution featuring very large susceptance magnitudes. Anders[1] has derived the following expression for these distances that will ensure a solution of minimum size provided that the N successive frequencies to be matched are separated by about the same interval:

$$D \text{ (in } \lambda \text{ at } f_{mean}) = \frac{Q.f_{mean}}{\Delta f} \frac{N-1}{2N} \quad (2)$$

where N is the number of frequencies to be matched simultaneously

Δf is the difference between the highest and lowest frequency

f_{mean} is the mean of the N frequencies

Q is any odd integer

D is the distance between successive mismatches.

The distances given by this formula should be regarded as the ideal; the nearest distances implementable in practice may still give a usable result.

In the finally computed solution, each susceptance will be placed within $\pm \lambda_{mean}/4$ of the position entered intially.

Most practical RF feeders are coaxial in construction and it is convenient to achieve an increase in susceptance at a specific point by means of a metallic sleeve on the inner conductor.

The effect of such a sleeve may be computed by treating it as a short section of line of lower Z_o, and the following approximate formula for the reflection co-efficient of a capacitive sleeve on a matched line may be used:

$$|\rho| \simeq \frac{\pi L}{\lambda}\left(\frac{Z_o}{Z_o'}, -\frac{Z_o'}{Z_o}\right)$$

where L is the length of the sleeve (in units consistent with λ)

Z_o is the line characteristic impedance.

Z_o' is the characteristic impedance of the sleeve.

In admittance co-ordinates, the angle of ρ_i referred to a point mid-way along the length of the sleeve is $+90°$ for small values of $|\rho|$.

Note that:-

(i) A relatively long sleeve of small diameter is equivalent to a shorter one of larger diameter; the former may be preferred if voltage breakdown is an important consideration.

(ii) It is perfectly feasible to place a shunt inductance (angle of $\rho_i = -90°$) across the line by reducing the diameter of the inner over a limited length, although this arrangement is not convenient in practice, because the position of such an undercut cannot easily be altered to allow for adjustment.

(iii) Although the reflection co-efficient of a sleeve as computed by equation (3) will be proportional to frequency, the behaviour of a practical sleeve deviates from this simple model because of the finite end surfaces, which set up fringing fields. This 'end-effect' is more significant at UHF where sleeve lengths are relatively short.

BANDWIDTH CONSIDERATION

It is often necessary to improve the impedance bandwidth of aerial systems on one or more channels. Consider the reflection coefficient plot shown in fig.2a. This is an actual measured plot referred to the input port of an unmatched UHF aerial array with a distribution feeder length of about 13 metres and represents the match of the system at frequencies spanning UHF Channel 53.

If such an aerial array is to be used to radiate a television signal, then its input match must, typically, not exceed the specification limit shown in fig.3 in order to avoid degradation of the picture by delayed image radiation. It is well established that any system of matching applied near to the input port will not affect the inherent bandwidth of the system, so that if, for example, the vision carrier reflection co-efficient is minimised, then the plot depicted in fig.2(a) will be shifted to place f_v at the chart centre without altering its basic shape to any extent: in this case it is clear that the resulting input match would exceed the specification limit set by fig.3.

The first step in improving bandwidth should always be to refer measured plots to an accessible position near to the source of the principal mismatch. In the case of this example, the radiating

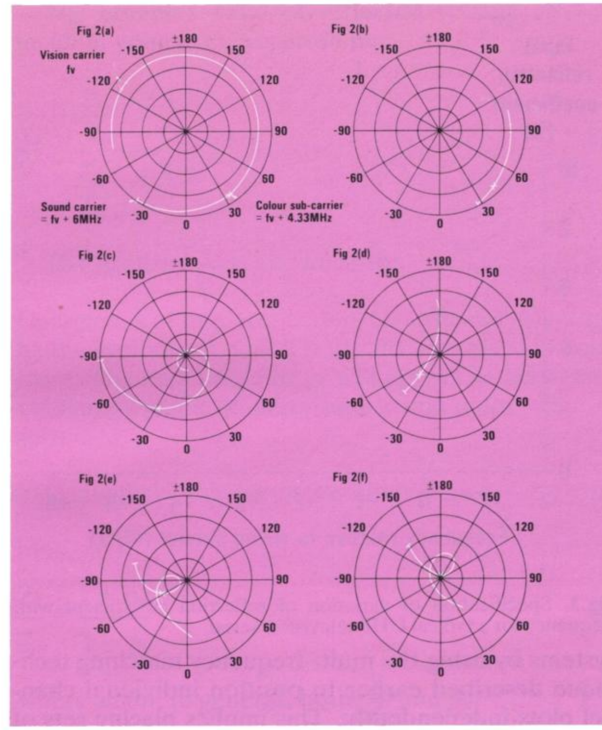


Fig.2. Measured impedance of a UHF aerial system at channel 53, distribution feeder length 13m.

Fig.2a. At the input port of the unmatched array.

Fig.2b. Referred to a position near to the radiating panels.

Fig.2c. With vision carrier moved to chart centre.

Fig.2d. Positioned to reduce the spread of the plot at the aerial input.

Fig.2e. Figure 2d referred to the input port showing improvement in bandwidth over figure 2a.

Fig.2f. Figure 2e with additional sleeve near the input port to control the plot.

elements are the main source of error, and if the measurements at the input port represented by fig.2(a) are referred to a datum position near to the radiating panels then the plot shown in fig.2(b) will be obtained. This displays a much improved bandwidth and although a better overall match at the input would result by moving the vision carrier point to the chart centre, [see fig.2(c)], it is better to position the plot as shown in fig.2(d). A single sleeve near to the datum position will achieve this. If this plot is now referred to the aerial input, it will be seen that the bandwidth has improved due to the compensating action of the line [see fig.2(e)]. A single sleeve may now be applied near to the input port of the system to achieve the plot shown in fig.2(f) and this easily meets the specification limit of fig.3.

Broadbanding can be extended to multi-channel

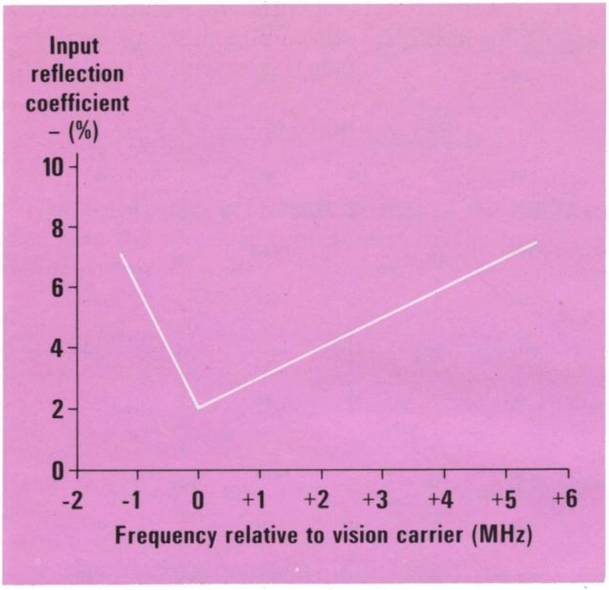


Fig.3. Specification of variation of reflection co-efficient with frequency for a typical UHF television aerial.

systems by using the multi-frequency matching technique described earlier to position individual channel plots independently. This implies placing sets of sleeves on the distribution feeder both at the datum and input port positions.

All compensation techniques that rely on differential reference through transmission lines fail if the line involved is too long. In the present application, this implies that matching elements must be accommodated within the first $3\lambda_d/8$ of feeder, where λ_d is the wavelength computed at the difference frequency equal to the bandwidth of an individual channel.

Conversely, if the line is too short, very large offsets will be needed to obtain worthwhile compensation, and it is usually more satisfactory in these cases to optimise the plot, centering it as near as possible to the source of the mismatch.

Practical Aspects

Reflection co-efficient is perhaps the most important parameter requiring measurement at high frequencies, and its accurate determination in a coaxial system is best accomplished by the use of a reflectometer which mates with the particular size of feeder in use. Provided that the directivity of both probes in the reflectometer is sufficiently high >40dB is a typical specification), errors will be small. Where adaptors of poor match are used to interface a reflectometer of the wrong size, or where the directivity of the reflectometer is poor, then local offset errors will occur. Figure 4(a) shows a load mismatch measured through a long feeder where local offset is present.

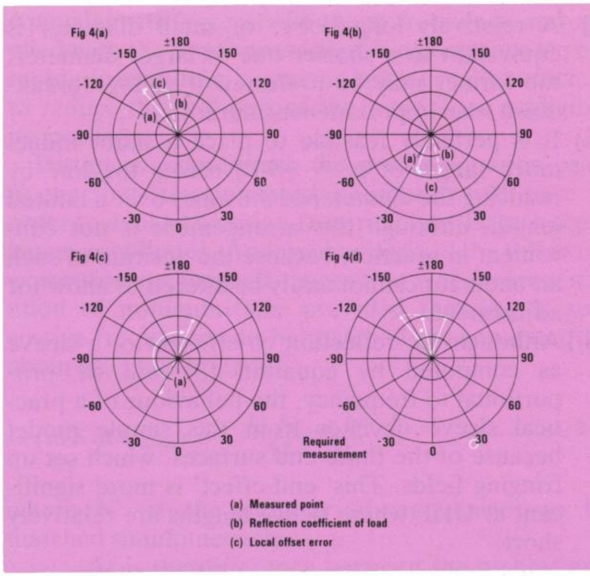


Fig.4. Reduction of measurement errors by 'quarter waving'.
Fig.4a. Measured reflection co-efficient in the presence of a local offset error.
Fig.4b. Reflection co-efficient when feeder length is extended by one quarter wavelength.
Fig.4c. Measurement of fig.4b referred back through one quarter wavelength.
Fig.4d. Averaging figs.4a and 4c removes the effect of the local error.

If the feeder is extended by $\lambda/4$ at the mean frequency and the measurement repeated, then the plot shown in fig.4(b) will be obtained which transforms to fig.4(c) after referring back through the quarterwave line. It is clear that averaging the plots shown in figs.4(a) and 4(c) will remove the local error [fig.(4d)]. This technique, known as quarterwaving, can be used to obtain a more accurate system measurement under conditions where it is not easy to separate measurement errors by inspection, although it is only effective if measurements and errors are small enough for linearity to apply.

A characteristic of all practical measurements made on transmission lines that are terminated in passive components is that the plot of reflection coefficient rotates clockwise with increasing frequency because the phase delay associated with the reflected signals increases with frequency. In certain cases, when a reflection co-efficient plot is referred through a large distance towards the load, it will be

found to exhibit an anticlockwise curvature with increasing frequency: such a plot is diagnostic of having been referred beyond the principal mismatch and does not correspond to a measurable reality.

Many passive systems, such as aerial systems, employ branch feeder networks which divide via matched transformers towards the load. Such systems can be regarded as one line for matching purposes over the operational frequency range of the transformers provided that, where appropriate, susceptances are placed at identical distances from the junction point on all branch feeders.

It is frequently necessary, after initially measuring a system, to refer the reflection co-efficient values obtained to a specified physical datum position, which may be some considerable distance towards the load from the point of measurement. There are several ways of achieving this, but the one most useful at UHF involves measuring the reflection co-efficient ρ_i of the system before and after the placement of a capacitive sleeve on the transmission line inner at the required datum position. A vector subtraction of corresponding 'before' and 'after' measurements will yield a set of vectors representing the effect of the sleeve alone. Provided that all reflection levels are less than about 0.15, then the vector change, referred to the position of the sleeve, will have an angle of approximately +90° on a reflection co-efficient chart, and this will apply at all frequencies. Hence the angular difference between the direction of any vector in the measuring plane and the +90° direction will be the referring angle appropriate to that particular frequency.

References

1.
M B Anders, 'Impedance Correction of Multi Channel UHF Aerials' *IBC 70 IEE Conference Publication No. 69*, pp. 61-63.

APPENDIX

Transmission Line Theory Relating to Multi Channel Impedance Matching

Practical uniform R.F. Transmission lines, terminated at a finite distance in a generalised impedance Z_t will reflect a part of the forward travelling wave unless $Z_t = Z_o$ (where Z_o is the Characteristic Impedance of the line).

The complex ratio of reflected voltage (V_r) to incident voltage (V_i) is given by;

$$\frac{V_r}{V_i} = \frac{\frac{Z_t}{Z_o} - 1}{\frac{Z_t}{Z_o} + 1} \tag{1}$$

The quantity $\frac{V_r}{V_i}$ is known as the Complex Voltage Reflection Co-efficient (denoted by the symbol ρ_v), and it is usually expressed in polar form. The magnitude of ρ_v will always lie between a maximum value of unity and a minimum value of zero.

Conversely, the normalised impedance $\frac{Z_t}{Z_o}$ is given by:

$$\frac{Z_t}{Z_o} = \frac{1 + \rho_v}{1 - \rho_v} \tag{2}$$

where again, in general, both, $\frac{Z_t}{Z_o}$ and ρ_v are complex quantities. If ρ_v in (2) above is purely *real*, then $\frac{Z_t}{Z_o}$ will be numerically equal to the VSWR on the line. Hence, in general:

$$\text{VSWR} = \frac{1 + |\rho_v|}{1 - |\rho_v|} \tag{3}$$

Voltage maxima occur at positions of current minima, and vice-versa.

If it is desired to work in admittance (Y) rather than impedance (Z), then:

$$\frac{I_r}{I_i} = \frac{\frac{Y_t}{Y_o} - 1}{\frac{Y_t}{Y_o} + 1}$$

Where $\frac{Y_t}{Y_o}$ is the terminating admittance normalised to the characteristic line admittance Y_o. (In this case I_r/I_i represents the Complex Current reflection co-efficient ρ_i).

It follows that:

$$\frac{Y_t}{Y_o} = \frac{1 + \rho_i}{1 - \rho_i} \tag{5}$$

Note that $\rho_v = -\rho_i$ (6)

Voltage and current reflection co-efficients can be inter-converted by changing the phase angle of either ρ_v or ρ_i by 180°.

If a load of reflection co-efficient ρ is connected to the end of a lossless transmission line of length 'l' then the magnitude of ρ measured at the input of the line remains unchanged, but its phase angle shifts by twice the electrical length of the line at the frequency concerned, i.e.,

$$\rho \text{ (line input)} = \rho \text{ (load)} \times e^{-j2\theta} \tag{7}$$

where θ is the electrical length of the line at the frequency of measurement

(i.e., $\theta = \frac{2\pi l}{\lambda}$)

This process of transforming reflection co-efficient from one point on a line to another point is known as 'referring'. When referring towards the generator, the angle of reflection co-efficient becomes more negative and vice versa. It is clear that the normalised input admittance (or impedance) of a mismatched line, (as given by equations (5) or (2) respectively), is a function of the electrical length of the line for any given load mismatch.

Elementary Matching

Most simple matching problems can be solved using the above relationships bearing in mind that:

(i) The total voltage on a mismatched transmission line is given by the sum of the forward and reflected voltage wave components, i.e.;

$V_t = V_i + V_r$ and $\frac{V_r}{V_i} = \rho_v$ hence:

$$V_t = V_i\,(1 + \rho_v) \tag{8}$$

Similarly, $I_t = I_i + I_r$ and $\frac{I_r}{I_i} = \rho_i$ hence:

$$I_t = I_i\,(1 + \rho_i) \tag{9}$$

(ii) Forward and reflected waves will suffer attenuation by a factor of α per unit length, due to dissipative feeder losses: this requires that the magnitude of ρ be reduced by a factor of $(2\alpha l)$ when referring through a line of length l.

Note that dissipative losses in transmission lines arise because of:

(i) The finite series resistance of line conductors.
(ii) Imperfections in the dielectric insulation between conductors.

At low frequencies, resistive components in the line equivalent circuit become comparable in magnitude with the reactance offered by the inductive and capacitive elements, and as a consequence attenuation, phase velocity and Z_o all vary with frequency. (In addition, Z_o evaluates as a complex quantity.)

At RF, lines are, in general, non-dispersive implying a purely resistive Z_o together with an attenuation characteristic dependent only (to a first order) on series conductor resistance. (This latter increases roughly in proportion to the square root of frequency as a consequence of skin effect).

At very high frequencies, dielectric loss may additionally contribute to the overall dissipation.

At any single frequency, a mis-matched load will give rise to a reflection co-efficient on the lossless line which at certain positions is representative of purely reactive error. These positions, spaced every $\lambda/2$ along the line for each case, occur when:

$$\text{Arg}\,(\rho) = -90 + \sin^{-1}|\rho| \tag{10}$$

$$\text{Arg}\,(\rho) = +90 - \sin^{-1}|\rho| \tag{11}$$

The magnitude of the reactance or susceptance found at these positions (in normalised units) is given by;

$$S = 2 \times \tan\,[(\sin^{-1}|\rho|)]$$

Conversely, a pure inductance or capacitance placed on a lossless matched line will give rise to a reflection co-efficient of magnitude:

$$|\rho| = \sin\,[\tan^{-1}(S/2)] \tag{12}$$

If $\rho = \rho_{v'}$ then S represents series connected *Reactance* in normalised units

If $\rho = \rho_{i'}$ then S represents shunt connected *Susceptance* in normalised units

Hence a mismatched load on a lossless line may be corrected at any one frequency by placing a single reactive mismatch on the line at an appropriate position.

Any such simple correction process may be thought of purely in reflection terms: the load reflec-

tion co-efficient and that of the mismatch are of equal magnitude, but the mismatch is so positioned that its reflected wave cancels with that of the load. Note, however, that at the position of the correcting mismatch:

(i) Forward travelling wave components are reduced in amplitude slightly because a fraction of the total energy is reflected backwards.

(ii) A percentage of the backward travelling wave component, (representing energy reflected from the load), is re-reflected forward.

The foregoing implies the establishment of multiple reflections between the load and the correcting mismatch which increases the group delay of the line as a transmission system and gives rise to the possibility of voltage/current amplification.

Hence, it can be shown that for a lossless line system;

$$V_{max} = (\text{VSWR} \times \text{nett power transmitted} \times Z_o)^{1/2} \qquad (13)$$

where V_{max} =
RMS value of the peak of the voltage standing wave

and

$$I_{max} = \left(\frac{\text{VSWR} \times \text{nett power transmitted}}{Z_o}\right)^{1/2} \qquad (14)$$

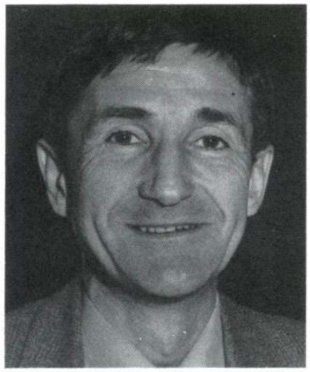
where I_{max} = RMS value of the peak of the current standing wave.

Note that simple matching procedures may be applied to RF systems where line loss is significant, provided that the magnitude of ρ is reduced by twice the 'one-way' line attenuation when referring towards the generator.

A transmission line will always dissipate the minimum of heat if it is operated without a reflected wave being present; if reflected waves are present, the overall line loss increases because of the heat dissipation caused by the reflected wave component. The factor by which 'one-way' attenuation must be multiplied in order to compute the total effective attenuation is given by:

$$\frac{1 - |\rho_{input}|^2}{1 - |\rho_{load}|^2} \qquad (15)$$

TED FORD joined the IBA in 1972 and is currently Head of Service Area Planning Section. Until recently he was Principal Engineer in the Masts and Aerials Group, where his responsibilities included the provision of new transmitting aerials and channel combining equipment for MF and VHF radio and UHF television. He obtained a BSc(Eng) degree in 1963 and before joining the IBA he worked for eight years on radar and spacecraft projects involving microwave and VHF aerials.



Rotamode Channel Combining Equipment

by E. T. Ford

Synopsis

At the present date (1985) there are 160 Marconi Rotamode Units of various types in service at 49 UHF main stations. A large proportion are channel combiners, installed in matched pairs to enable the UK's fourth television channel to be fed into existing multi-channel transmitting aerials at IBA and BBC main stations. In nearly all instances these combiners carry two channels only but in a unique arrangement at Bilsdale the equipment had to be installed at a location where it carries all four services.

The UHF Rotamode channel combiner comprises a pair of coupled cavities in cylindrical waveguide, containing input and output stripline coupling loops. Its method of operation is described in a simplified manner and the measured performance of the Bilsdale equipment is presented.

INTRODUCTION

As part of the implementation of the Fourth Television Channel, nearly all UHF Main Stations in the United Kingdom now incorporate some form of the Marconi 'Rotamode' equipment used in the following roles:

(i) Bandpass filters for Marconi passive-reserve transmitters at 11 stations.
(ii) Bandpass filters for Channel 21 transmitters at six stations.
(iii) Sound-Vision Combiners for Marconi transmitters at 25 stations.
(iv) Multi-channel combining equipment at 47 stations.

In the case of the multi-channel equipment the IBA awarded a contract to Marconi to supply and install Rotamodes for Channel Four at BBC as well as IBA Landlord stations and it is the channel-combining aspect that forms the subject of this article. The combiners are identical in their principles of operation to the bandpass filters, including the Channel 21 filters. Sound-Vision combiners, on the other hand, differ slightly in their operation[1].

ROTAMODE DIRECTIONAL FILTER

The Rotamode achieves its performance from a combination of two directional couplers and a waveguide bandpass filter that can support two orthogonal modes simultaneously. These modes are in time quadrature and produce a synchronously rotating circularly polarised field inside the waveguide, whence the name 'Rotamode'.

A single channel f1 may be added to a line carrying several other channels f2, f3, f4 etc, and there will be full isolation between the f1 transmitter and the other transmitters, in both directions.

A signal f1 starting at the top left-hand port (in fig.3) and travelling right is split by the left directional coupler into two equal components which are 90° out of phase as they enter the bandpass filters. If the two filters have identical attenuation and group delay characteristics in the passband the signals arriving at the right-hand directional coupler are recombined so as to pass out to the aerial at lower right. Only if there are inequalities in the two filters and imperfections in the two couplers will there be a small residual coupling of signal f1 into the wideband

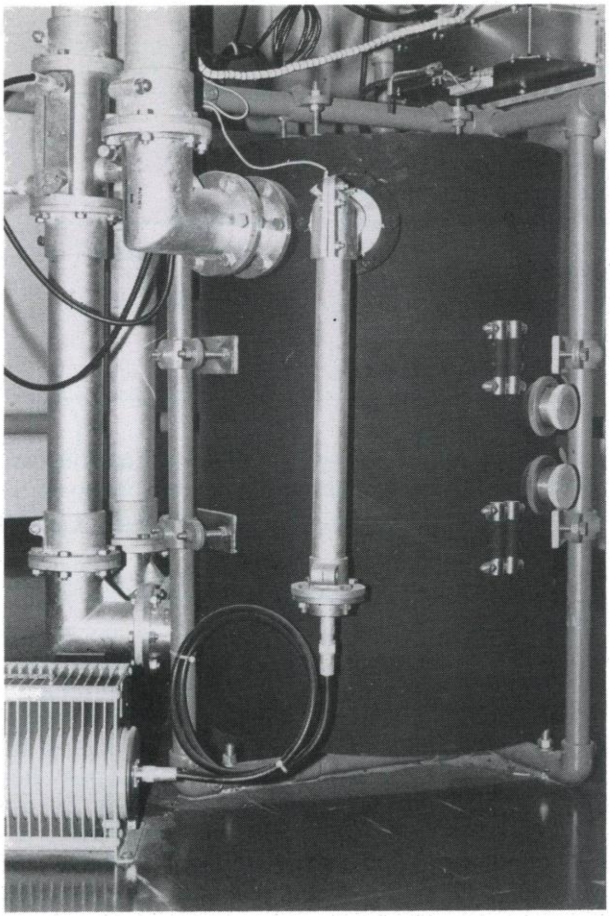


Fig.1. One of a pair of Rotamode channel combiners installed at Dover.

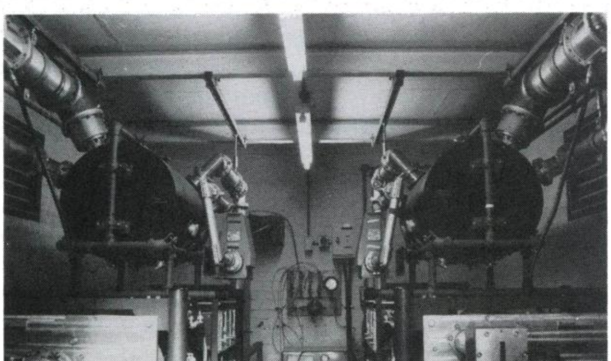


Fig.2. A pair of Rotamode combiners installed at Sandy Heath.

port at top right. In a Rotamode unit the two identical bandpass filter responses are in reality achieved by two signals with orthogonal modes in a single waveguide filter having circular symmetry.

Consider now the wideband input port at the top right. Signals entering and travelling left are split into two components of equal amplitude and in time quadrature by the action of the right-hand directional coupler. Provided that the frequencies differ sufficiently from f1, so that the filters present a reflection coefficient greater than 98%, (ie more than 14 dB down on the rejection skirts of the filter) the components will be reflected back into the right-hand coupler, where their phase relationship is such that they will be totally combined into the aerial output port.

The transmitters do not see the short circuit produced by the filters. They see a nominally wideband matched input port. However, imperfections in the right-hand coupler can produce residual mismatches into this port.

For the wideband port to achieve the desired characteristics it is essential that the filters be of the reflection type. There is an important distinction between 'reflection-type' and 'absorption-type' bandpass filters that needs emphasising: an absorption filter directs stopband energy away from the main signal path through the unit and dissipates it in resistive loads, resulting in a so-called 'constant impedance' characteristic, which means that the filter is a 50-ohm device at all stopband and passband frequencies. A reflection filter on the other hand contains no absorptive components and can only produce its stopband attenuation characteristic by reflecting energy back to the transmitter. Its reflection co-efficient characteristic resembles its attenuation characteristic. Very large input reflection co-efficients (ρ) approaching unity are encountered in the stopband according to the simple relationship:

$$\text{Filter Insertion Loss, dB} = 10 \log_{10} (1-\rho^2) \ldots\ldots (1)$$

The equivalent-circuit filter needs to be a reflection-type in its stopband to allow the wideband input port to operate, and indeed a reflection filter is, in general, much easier to construct if a critically-shaped passband response is to be realised, as is needed in the Channel 21 Filter. Over the passband a Chebyshev equal ripple characteristic (0.25 dB max) is employed to obtain the required attenuation skirts in the stopband and to put attenuation zeroes close to the vision and sound carriers in the channel combiner. This results inescapably in substantial reflection co-efficient ripples in the passband. Equation 1 shows that an attenuation of 0.25dB results from a 24% reflection co-efficient and such a mismatch

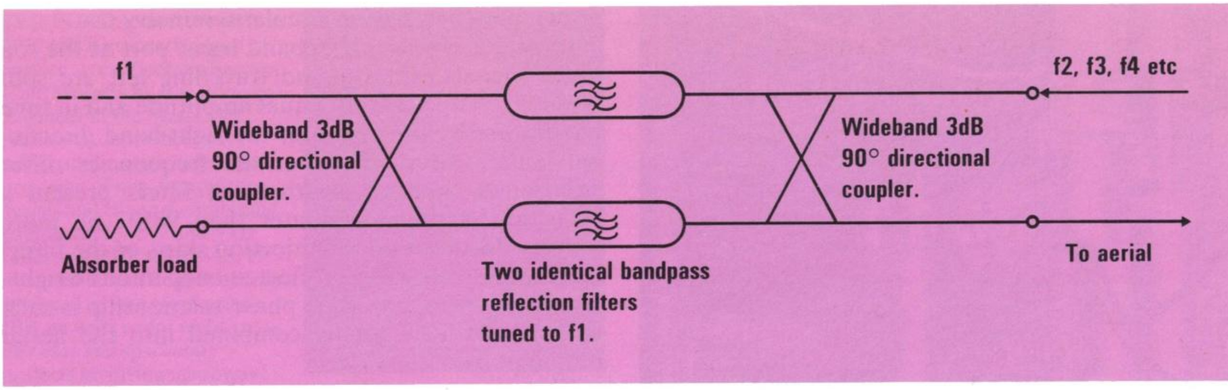


Fig.3. An equivalent circuit of a Rotamode used as a combining filter.

would be unacceptable for system performance. However, the Rotamode overcomes this problem by converting its f1 channel input port into an absorption-type filter by virtue of the left-hand directional coupler. Exactly analogous to the operation of the wideband port, the f1 port does not see the reflection co-efficient of the filters because all the reflected energy is directed into the absorber load.

The equivalent circuit in fig.3 represents a channel-combining unit, but can equally well represent a bandpass filter, such as the Channel 21 Filter, by disconnecting the transmitters from the wideband port and terminating it with a balancing load. The equipment becomes a two-port device with constant impedance characteristics, passing the f1 channel only between the top left and bottom right ports.

The equivalent circuit in fig.3 is valid for all considerations of signal paths associated with transmitter cross modulation, intermodulation, imperfections in components, reflections from interfaces and so forth. For example, it is seen that none of the transmitters will be isolated from mismatches generated by the aerial or any other external component. Obviously this must be true because the Rotamode is a four-port reciprocal device.

PHYSICAL REALISATION OF A ROTAMODE FILTER

The filter comprises a round waveguide transmission line divided into a number of cavities. An input directional coupling strip is fitted inside one end of the first cavity and a similar output coupling strip is located at the other end of the last cavity. In principle a single cavity could be used, or any even or odd number N. The bandpass response is simply that of an N-section filter. For a channel-combining application, where transmitters are spaced not less than three channels apart, only two cavities are needed. (A passive-reserve transmitter bandpass filter needs four cavities, and the Channel 21 Filter needs nine).

The derivation of a two-cavity Rotamode is shown in fig.4 in four stages:

(a) A direct-coupled resonator filter in any typical transmission line, comprising three shunt inductive susceptances suitably spaced to produce a bandpass filter response. The filter is a reflection type because no resistive components are introduced (typical values for a Chebyshev 0.25dB ripple response at Channel 23 are given), the susceptances and spacings being derived using Cohn's method[2].

(b) The equivalent filter in round waveguide, using circularly symmetric inductive irises comprising round holes in conducting plates.

(c) A stripline coupler in round waveguide which simultaneously provides electric coupling and magnetic coupling to produce two orthogonal field components of equal amplitude in time quadrature, equivalent to a single circularly-polarised field[1].

(d) The two-cavity Rotamode fully derived by blanking off the end irises and inserting two stripline couplers fed by 50-ohm coaxial lines. The direction of rotation of the field imparted by the input coupler is dependent upon which input port is fed and hence controls which output port obtains the signal.

For correct operation, each of the two orthogonal field components must experience exactly the same bandpass filter response – see the two component 'filters' shown in the equivalent circuit in fig.3 – and this would be the case if the waveguide cavities and irises were truly circularly symmetric and the di-

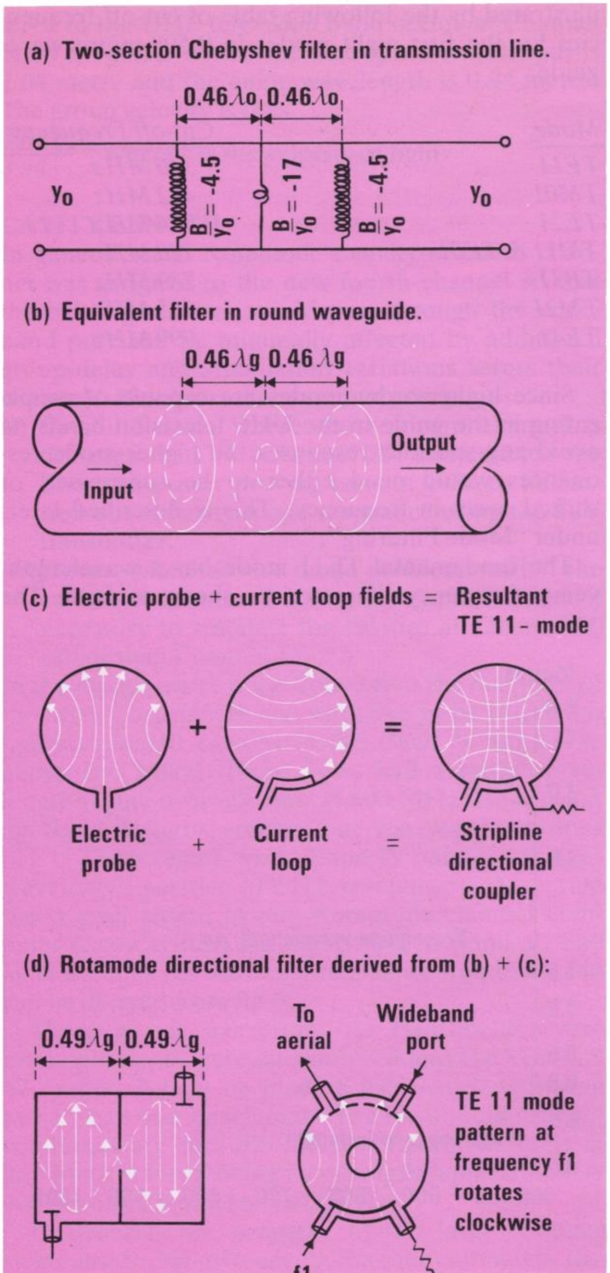


Fig.4. Derivation of the Rotamode filter from a transmission line filter (see text).

rectional couplers were perfect in their directivity. In a real unit, tuning slugs in the cavities are needed to correct for residual errors in manufacture. If any irregularities were to exist, for example a slight dent or ellipticity in cross section, the TE11 mode would become elliptically polarised. Elliptical polarisation is equivalent to the superimposition of a small unwanted component of anti-clockwise rotating field on to a wanted larger component of clockwise rotating field. The unwanted component excites the other ports of the directional input and output couplers, giving rise to extra power into the absorber load, together with undesired coupling between the two transmitter ports.

Symmetry in two orthogonal planes implies that a Rotamode filter could equally well be constructed from square waveguide, but circular was chosen by Marconi for manufacturing reasons.

CONFIGURATIONS FOR MULTICHANNEL COMBINING

A Rotamode combiner cannot operate in a '2 + 2 channel' configuration. It is equivalent to an 'N + 1 channel' unit, where N is 1, 2, 3 or more. The fourth channel is combined with ITV – or BBC2 in a few instances – by introducing a '1 + 1 channel' Rotamode combiner between the transmitters and the existing Lorentz Ring '2 + 2 channel' unit as in fig.5a.

To illustrate yet another possible location for a Rotamode, see fig.5b. Where an existing Lorentz Ring combining system is only tuned for three channels, which occurs in the "'2 + 1' unit at Bilsdale," a '3 + 1' Rotamode has been introduced between the Lorentz Ring system output and the half-aerial feeder so that the fourth channel is combined at this junction. The wideband port of the Rotamode then has to be matched to, and carry the power of, three channels rather than one.

This unique four-channel arrangement at Bilsdale results in the highest mean power (48 kW total) and highest peak voltage (7 kV) carried by the output coupler of a Rotamode channel combiner at any UHF Main Station and moreover leads to the widest bandwidth requirement.

PROPERTIES OF CIRCULAR WAVEGUIDE

The group velocity of any packet of energy travelling along a waveguide is frequency-dependent. It approaches the free-space velocity at high frequencies but becomes smaller at low frequencies, tending to zero at the cut-off frequency.

At a given frequency the guide wavelength (λg), depending as it does on phase velocity rather than group velocity, is actually larger than the free-space wavelength (λo):-

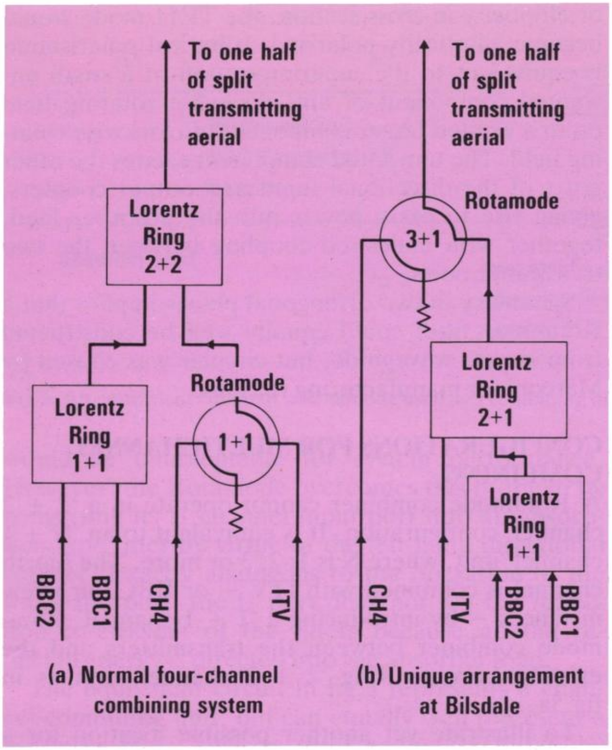


Fig.5. Configurations for multi-channel combining
(a) Rotamode filter as used in a normal four channel combining system.
(b) The unique arrangement required at Bilsdale to combine Channel 4 with the other three channels.

$$\lambda_g = \frac{\lambda_o \lambda_c}{\lambda(\lambda_c^2 - \lambda_o^2)^{1/2}} \text{ where } \lambda_c = \text{cut-off wavelength}$$

The cut-off wavelength determines the lowest frequency that can propagate in any waveguide. In a circular waveguide it is related to the diameter D:-

$D = 0.586\,\lambda c$ (see Ref.3)

It is a feature of waveguide that its diameter has to be an appreciable fraction of the free-space wavelength, which explains why it is usually encountered only at UHF frequencies and above. Marconi Ltd has chosen a waveguide internal diameter of 0.508 metre (20 inches) to be a size suitable for all Rotamodes from UHF Channel 21 up to Channel 68, there being no change in diameter between Band IV and Band V. The cut-off frequency is 346 MHz.

All Rotamodes use the fundamental TE11 mode, (TE = Transverse Electric), which has a longitudinal magnetic component. Higher modes of the TE and TM type (TM = Transverse Magnetic) can exist, as illustrated by the following table of cut-off frequencies for the first eight modes in 0.508-metre waveguide:

Mode	*Cut-off Frequency*
TE11	*346 MHz*
TM01	*452 MHz*
TE21	*574 MHz*
TM11 & TE01	*720 MHz*
TE31	*789 MHz*
TM21	*965 MHz*
TE41	*999 MHz*

Since higher-order modes are capable of propagating in the guide in the UHF television bands, to avoid any spurious responses the higher-mode resonances would need either to be suppressed or shifted away in frequency. This is described later, under "Mode Filtering".

The fundamental TE11 mode has a wavelength-versus-frequency characteristic shown in fig.6. The

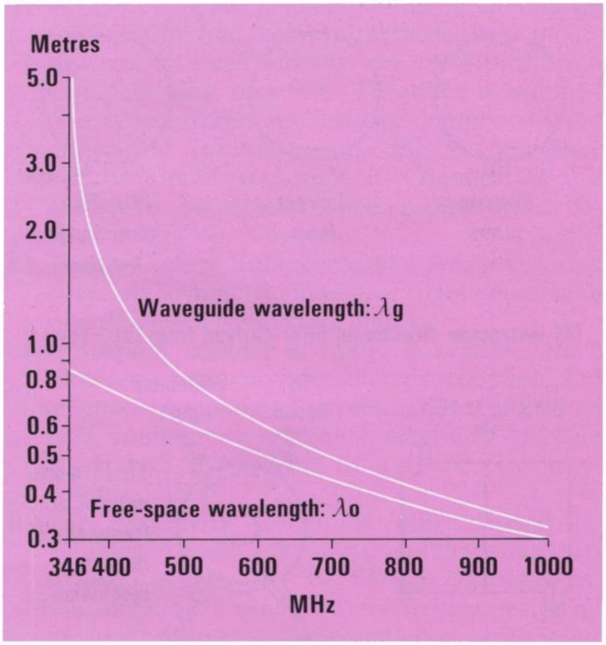


Fig.6. Graph showing the variations of waveguide wavelength with frequency for TE11 mode.

waveguide wavelength is asymptotic to the free-space wavelength at high frequencies but becomes much larger and tends to infinity as the cut-off frequency of 346 MHz is approached. At cut-off the group velocity tends to zero. The longest guide wavelength and hence lowest group velocity encoun-

tered in the UHF television band occurs at Channel 21 (470 MHz), where the free-space wavelength is 0.64 metre and the guide wavelength is 0.94 metre. The group velocity is thus

$\frac{0.64\,c,}{0.94}$ where c is the velocity of light.

CAVITY LENGTH

In general, each Rotamode's cavity-resonant channel was assigned to the new fourth channel so that the existing services would pass through the wideband port and be minimally affected by additional group-delay and attenuation variations across their channels. However, there are two exceptions:

(i) Crystal Palace, where the resonant channel is designed for ITV (Channel 23) and enabled the first high-power proving tests in 1980 to be carried out at cavity-mode resonance using the ITV transmitter.

(ii) Dover, where severe mode problems within the cavities encountered during development made it necessary to resonate the cavities at Channel 66 rather than Channel 53.

In the design phase it was found that the diameter of the input-port flanges was too large to be fitted into half-wavelength cavities in the Band V units (i.e. above 600 MHz). From fig.6, half a guide wavelength is just over 300 mm at 600 MHz. As a result all Band V units employ one-wavelength cavities (TE112 resonance), whilst Band IV units retain half-wavelength cavities (TE111 resonance). As a rule the overall length of any Rotamode channel combiner (two cavities) is roughly $2\lambda g$ in Band V, and λg in Band IV, at the cavity resonance frequency and can be derived from fig.6.

There is one exception. At Heathfield severe mode problems were encountered and the Channel 67 cavities had to be made a half-wavelength long rather than one wavelength. The Heathfield channel combiners are thus the smallest Rotamode units at any Main Station, being λg in length (approximately 400 mm) at Channel 67 (842 MHz).

In general, the presence of the large coupling strips inside the two-cavity channel combiners disturbs the orthogonality between the TE11-mode field components, and compensation from the cavity tuning slugs is necessary during setting-up. For this reason, four slugs are provided in each cavity.

THE STRIPLINE COUPLERS

The length of each coupling strip is approximately $\lambda_o/4$ at the cavity-resonance frequency in order to achieve maximum coupling into the cavity.

At frequencies away from cavity resonance, the output coupling strip provides the direct wideband connection to the aerial port (see fig.4d). Here the stripline is acting virtually as a short link between two 50-ohm coaxial ports and ideally should have a characteristic impedance of 50 ohms. Thus the width of the strip has to be about five times its separation from the cavity wall.

At cavity-resonance frequency, the coupling factor into a Rotamode filter's outermost cavities is dictated by the magnitude of the outermost pair of susceptances in the equivalent transmission-line circuit design for the Chebyshev bandpass response (e.g. see fig.4a). The loaded Q of the cavity is related to the normalised susceptance thus:

$$Q \simeq \left[\frac{B}{Y_o}\right]^2 \cdot \frac{\pi\lambda_g^2}{2\lambda_o^2}$$

In the channel-combining case, where only two cavities are needed, the normalised susceptance is in the range 4 to 8, leading to loaded Q's of 60 to 115. The coupling factor is dependent on the location of the strip as measured along the axis of the cavity from the flat end-wall, being minimal near the wall and a maximum at $\lambda_g/4$ from the wall or from the central iris, ie. at an E-field maximum. In reality this location is constrained because of two major practical considerations:

(i) In half-wavelength cavities the couplers can not be fitted at the E-field maximum because the cavities are factory-assembled by joining two halves circumferentially at this point, where waveguide wall currents are largely circumferential, thus minimising longitudinal current flow across the join; (a colour-coded diagram of the E- and H-fields and waveguide wall currents for the TE11 (H11) mode is given in Ref.3).

(ii) The strips are wide and if placed close to the end walls spurious capacitive coupling would occur.

However, the voltage coupling factor is almost linearly proportional to the penetration (i.e. the spacing of the strip away from the inside surface of the cavity) for small penetrations. This adjustment is used to control the loaded Q, which is empirically determined by sealing off the cavity electrically, fitting it with a matched coupling strip, and inserting a small test probe to measure the 3dB bandwidth and hence the loaded Q of the cavity response.

In practice, the stripline couplers cannot be de-

signed to give a perfect 50-ohm match, so capacitive impedance-compensating sleeves are soldered around the inner conductors of the coaxial feeders connected to the couplers. These need to be placed close to the striplines in order to maintain a broad bandwidth and often some are fitted directly adjacent to the strips (see fig.7), whilst others are placed a short distance away in the input and output coaxial lines outside the Rotamode cavities. The sizes and locations of the sleeves are determined empirically to achieve an input reflection coefficient of less than 5% over each channel. The short lengths of coaxial line containing these sleeves thus form an integral part of the overall Rotamode system.

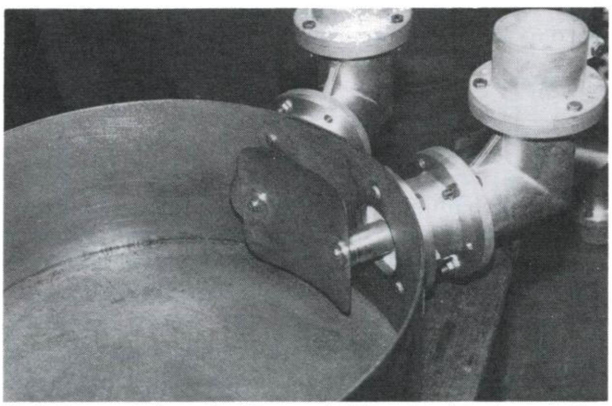


Fig.7. Internal view of a Rotamode filter showing the coupling strip.

INSERTION LOSS, HEAT DISSIPATION & THERMAL STABILITY

Owing to the large internal surface area of the waveguide cavities and the absence of a coaxial inner conductor the Rotamode channel combiners have acceptably low copper losses and heat dissipation at the transmitter powers used by the IBA and the BBC. With their outer bodies painted matt-black, all the channel combiners operate with moderate surface temperatures without any thermal control or forced-air cooling.

The units are constructed from copper-plated mild steel. Thermal expansion of 11 parts per million per °C causes the cavity diameters to increase as well as their lengths, the nett result being a frequency coefficient of −11 parts per million per °C (ie −5 kHz per °C at 470 MHz and −9 kHz per °C at 840 MHz). Such changes are insignificant in the UHF channel-combining application.

MODE FILTERING

Although, apart from their length, all Rotamode channel combiners are superficially similar in external appearance they contain internal mode-filtering components which differ markedly from one channel group to another.

Higher-order modes produced many unwanted resonances in the waveguide cavities during development. The Bilsdale combiners illustrate an example of this where, over the band 440-580 MHz, three cavity resonances were predicted:

TM010 at 452 MHz (not a problem)
TE111 at 490 MHz (the "wanted" resonance)
TM011 at 569 MHz (unwanted resonance in channel 33)

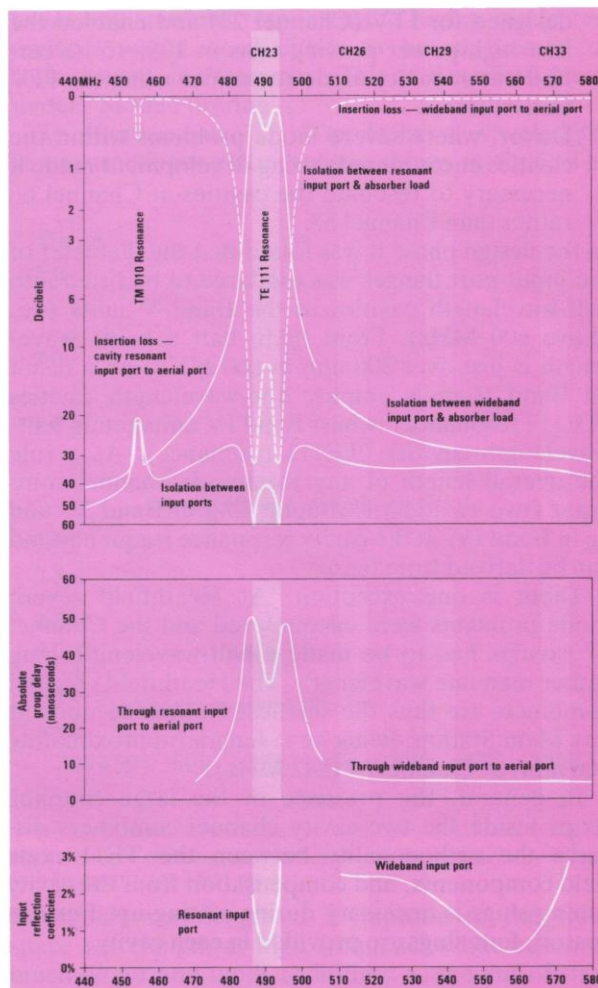


Fig.8. Results of factory test measurements made of the Bilsdale channel combining Rotamode filter.

To obtain a clean performance over channels 23, 26, 29 and 33 the TM011 resonance needed to be suppressed or re-tuned. Various forms of mode filter can be used to fully or partly short-circuit the E-field (e.g. longitudinal rods & loops for TM modes; lateral rods & loops for TE modes). In the Bilsdale cavities small square longitudinal loops were fitted to re-tune the TM011 resonance away to 615 MHz. The final factory test measurement of one of the pair of Bilsdale combiners (fig.8) shows no spurious resonance remaining within the band covering the four operational channels.

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Acknowledgements
Acknowledgements are extended to R. G. Wills, R. Hutchinson and D. S. Rose of Marconi Communication Systems Limited, for their technical advice during the preparation of this article.

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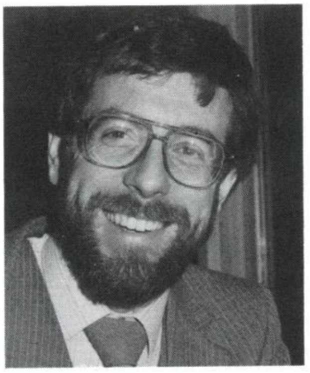

VHF RADIO TRANS-MITTING AERIALS

by D. R. Brian

Synopsis

The need to provide versatile and economic VHF transmitting aerials for Independent Local Radio (ILR) has presented the IBA and the aerial manufacturing industry with some unique problems. The variations in both the general and specific electrical and mechanical requirements have led to considerable development in recent years and this article describes the history behind the problems and outlines some of the solutions. Early systems used single-channel elements but further development to cover a new wideband requirement led to radiation pattern problems. The inclusion of slant polarisation as an option increased the choice of available aerials but recent developments in both directional and omnidirectional circularly polarised systems by the major UK manufacturers have further widened the scope for VHF radio transmitting aerial systems.

Introduction

At the start of ILR development in the early 1970s the IBA adopted a policy of using existing support structures, wherever possible, on which to mount its aerial systems. This meant attempting to obtain the required radiation characteristics from what was sometimes an imperfect location and resulted in the development of unique solutions within the electrical and mechanical constraints.

It was decided at the outset to specify circular polarisation and all aerials in the Phase 1 ILR programme (i.e. up to 1976) were built as narrow band, single channel systems with the exception of the dual channel aerial for the two London services. This led to the use of RCA 'BFG' elements as a suitable circularly polarised aerial at some sites e.g. London and Edinburgh, sometimes in a top-mounted cantilever location, giving good omnidirectional horizontal radiation pattern characteristics, and sometimes mounted on the side of the support structure resulting in a modification to the horizontal radiation pattern (hrp).

Crossed yagis were used in various configurations at other sites where a directional hrp was required in order to cover the service area. Further information on the performance of these systems, and a description of circular polarisation can be found in *IBA Technical Review 5* on Independent Local Radio and *Technical Review 14* on Sound Broadcasting.

Despite the economic and practical limitations imposed, reasonably good solutions were found to each problem, but generally each system will only operate satisfactorily over one channel.

Later Developments

At the start of Phase 2 of the ILR building programme in 1979, it was decided that, due to the

uncertainty of future frequency allocations within Band II, all VHF ILR aerials should be specified to be able to operate over the sub-bands 94.5 to 97.6 MHz and 100 to 105 MHz. This imposed yet another constraint on the aerial designer, added to which was the increasing difficulty with finding good locations at which to mount the aerials on suitable structures.

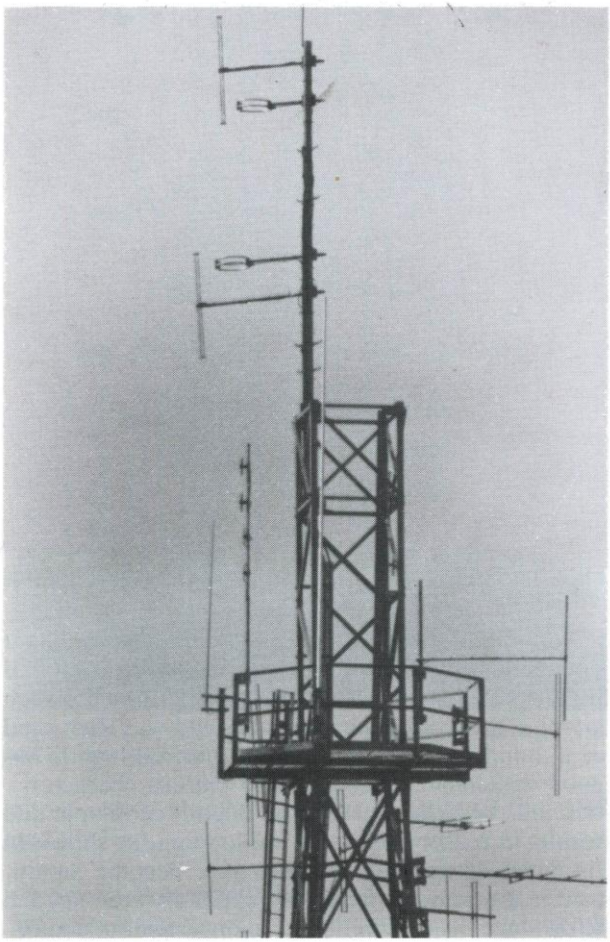

Fig. 1. C&S Antennas' loop and dipole circularly polarised aerial.

C&S Antennas Ltd had provided a loop and dipole circularly polarised aerial at some of the later Phase 1 sites (see fig.1). This consisted of a vertical dipole producing the vertically polarised component and a horizontal dipole, bent round into a loop, which produced the horizontally polarised component. The signals fed to the two elements had the required 90° phase difference for circular polarisation. This assembly was mounted on a pole and gave a reasonably good omnidirectional hrp (see

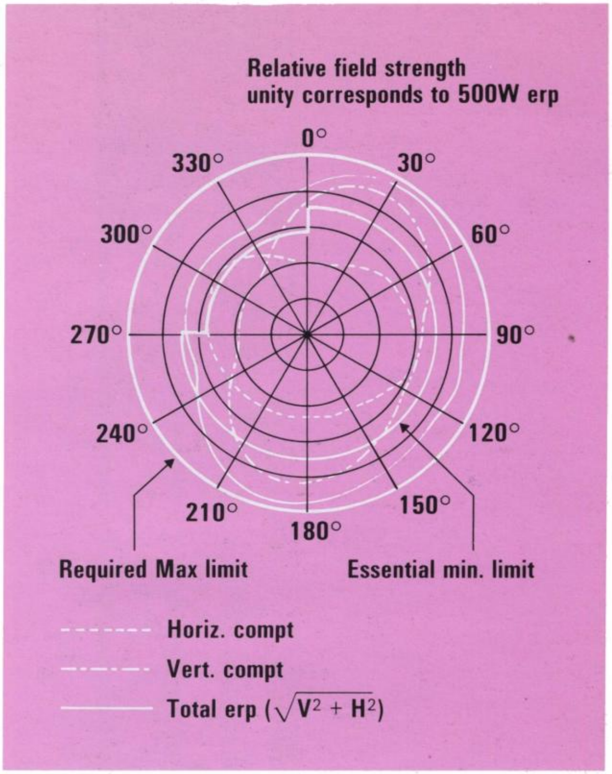

Fig. 2. Omnidirectional patterns achieved by C&S Antennas' loop and dipole aerial for Reading.

fig.2). A wideband version of this system was developed for Phase 2 of the building programme; however its hrp performance over the band was not fully satisfactory. The problem was especially noticeable when the aerial was further developed for use when mounted on the side of a support structure. Typically the 4ft square top of a UHF relay tower would be available. Figures 3 and 4 show a typical mechanical arrangement and hrp. A theoretical study of the loop and dipole system and its shortcomings appears elsewhere in this review.

However the system does represent an economic solution to the wideband, nominally omnidirectional, requirement and has been used at many sites. The alternative of using a combination of several directional elements, such as panels, mounted around a large structure, apart from being very expensive, would have imposed an unacceptable wind loading on the type of structures under consideration.

Circular v Slant Polarisation

Radiation from a nominally circularly polarised ae-

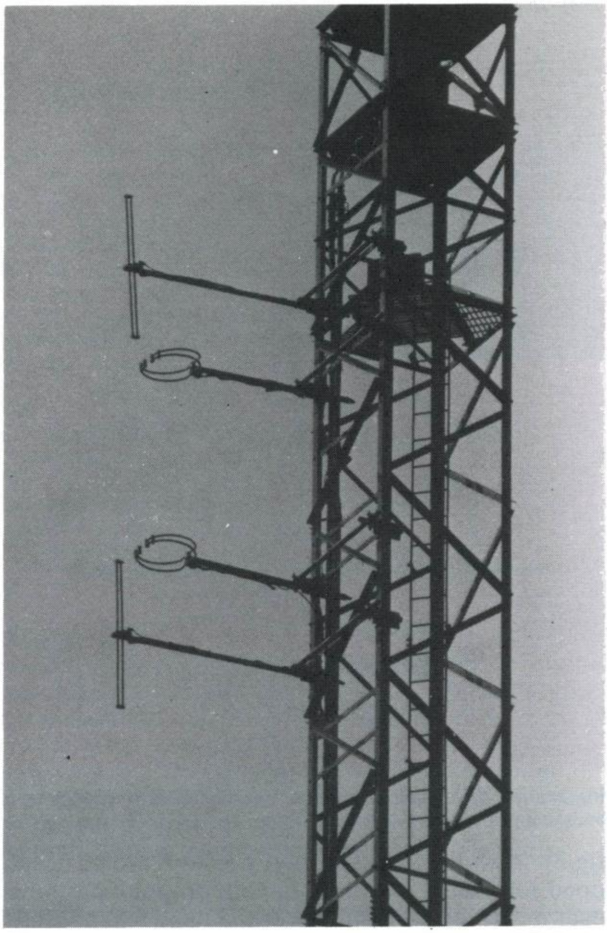

Fig.3. A C&S Antennas' loop and dipole aerial mounted on the Malvern UHF television relay tower for the Worcester ILR service.

rial is in reality elliptically polarised, and depends on the relationship between the amplitudes and phases of the horizontal and vertical components at any particular receiving location. The IBA provided circularly polarised aerials at all its Phase 1 sites but meanwhile systems were becoming available which produced mixed or slant polarisation. 'Mixed' covers all random horizontal and vertical component phase relationships and in the special co-phased case produces slant polarisation. The requirement that energy is radiated in both the horizontal and vertical planes is satisfied but there is still a plane at 90° to the resultant in which no energy is radiated. Circular, and to some extent random mixed polarisation, has the advantage that the resultant electric field vector rotates relative to a stationary observer, and hence energy is present in all planes.

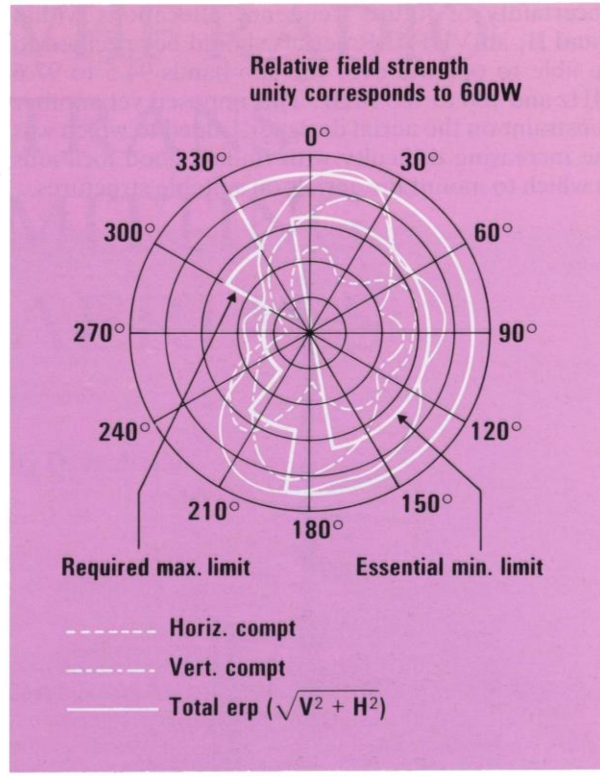

Fig.4. Radiation patterns for C&S Antennas' loop and dipole shown in fig.3.

Some BBC systems were engineered using a slant polarised skeleton slot radiating element developed by C&S Antennas. This was mounted on a screen giving a predictable performance and was then used as a 'building block' to form the desired hrp. It has good wideband impedance and pattern characteristics and the light construction is both economic and results in reasonably low wind loading, as shown in fig.5. However both factors could become significant in a situation where several panels were needed to produce the required pattern on a small structure.

With the availability of such attractive alternatives, the IBA modified its specification to allow the use of slant polarisation and employed the CSA panels for the Preston/Blackpool ILR aerial on the 9ft diameter steel cylindrical mast at Winter Hill. Figure 6 shows the original hrp achieved with two panels per tier compared to the desired template. It can be seen that an excellent fit was obtained and good wideband characteristics were maintained.

Meanwhile all three major UK aerial manufacturers were under pressure to offer equally versatile and economic systems to produce circular polaris-

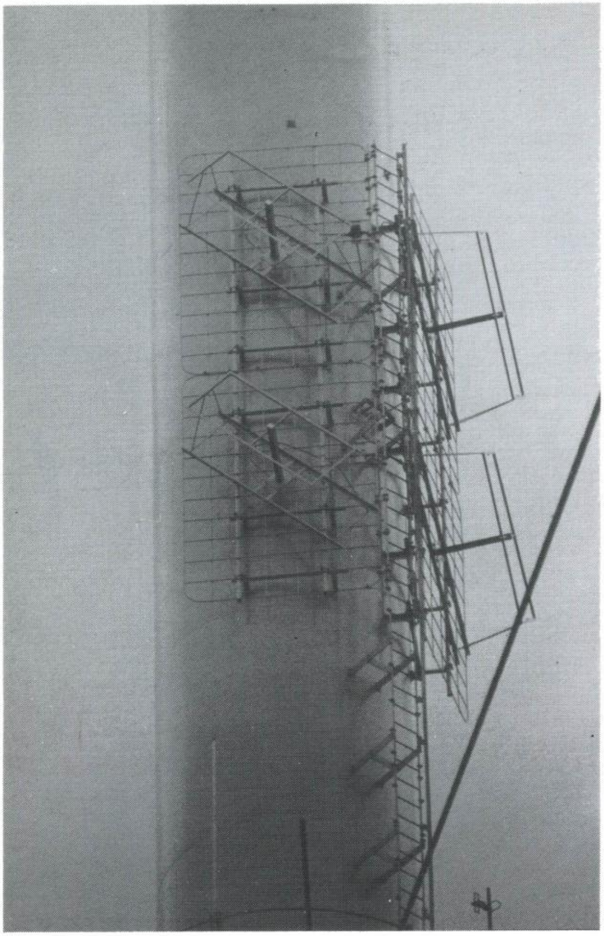

Fig.5. C&S Antennas' slant polarised skeleton slot aerial used for the Preston and Blackpool ILR service.

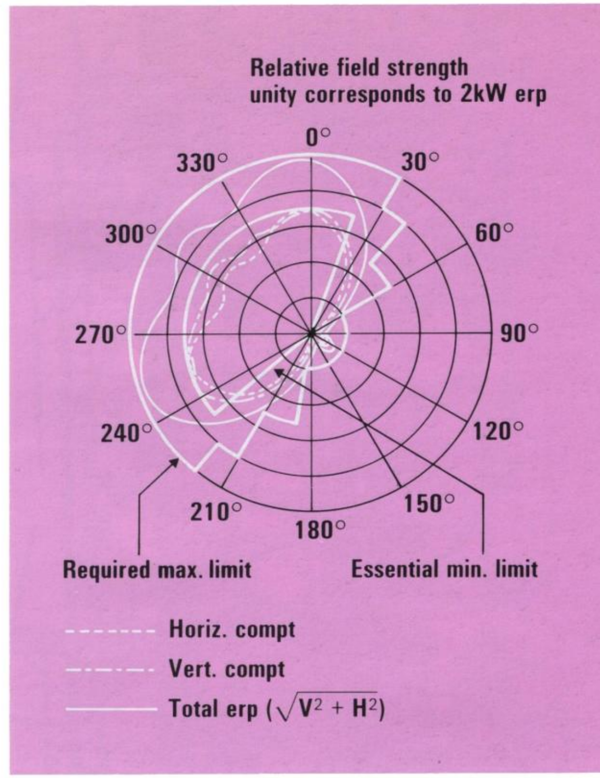

Fig.6. Radiation patterns achieved for the C&S Antennas' skeleton slot aerial (two panels per tier) shown in fig.5.

ation, which was now preferred by both the BBC and the IBA.

Circularly Polarised Panels

Manufacturers were encouraged to develop suitable circularly polarised panel aerials and offer them, when possible, to fulfil ILR specifications. It is to their credit that suitable panels were developed by all three contractors. To date the IBA uses Alan Dick & Co Ltd panels at Reigate and Sheffield, Marconi Communication Systems panels at Guildford and Brighton and C&S Antennas' circularly polarised panels at Dover. Photographs of the three types are shown in figs.7, 8 and 9. Each was designed to provide directional coverage of the service area and this is achieved by a combination of panel arrangement and variation in the power split. The patterns achieved in typical cases are shown in figs.10, 11 and 12.

The principle of operation of each type of radiating element is essentially the same, derived from the use of a pair of crossed dipoles mounted on a reflecting screen, each dipole being fed with equal power but with 90° phase difference. Two versions of each type of panel are available, one having a beamwidth suitable for combining with adjacent panels at 120°, e.g. on triangular structures, and the other having a narrower beamwidth suitable for mounting at 90° to adjacent panels. The 120° panels generally have 'arrowhead' dipoles to spread the beam whereas the 90° versions have straight dipoles, usually with parasitic side reflectors to narrow the beam. The dipoles are normally mounted at 45° to the horizontal and vertical planes to equalise the horizontal to vertical ratio. However this was found to vary at the edges of the band, especially in BBC applications where the specification covers the whole of Band II, 88-108 MHz. This problem was overcome in multi-tier arrays by reversing one of the crossed dipole elements on alternate tiers and

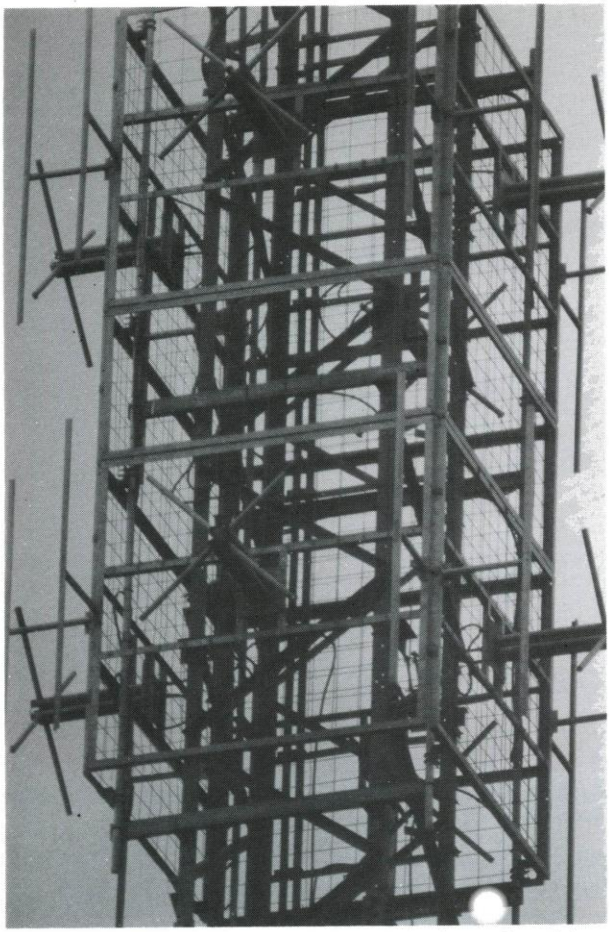

Fig.7. Alan Dick & Co circularly polarised panel aerial as installed at Sheffield.

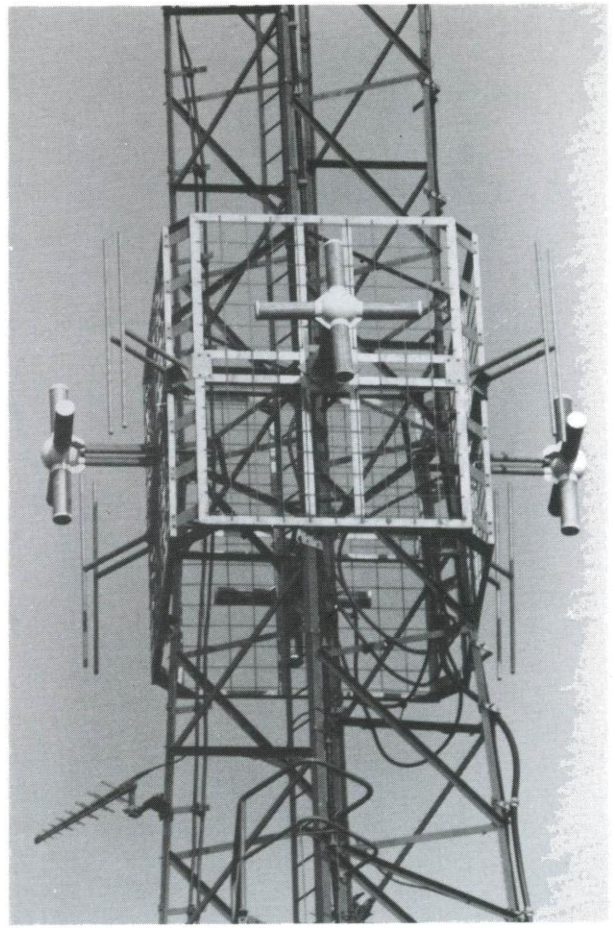

Fig.8. Marconi Communication Systems' circularly polarised panel aerial similar to that installed at Brighton.

adding appropriate lengths of feeder to compensate the phase in the distribution system. The alternative of mounting the dipoles in line with the vertical and horizontal planes, as used at Guildford, is not desirable in multi-tier arrays because of the differential coupling between horizontally and vertically polarised components of adjacent tiers.

Omnidirectional Aerials

In parallel with the development of panel aerials both Alan Dick and Co and C&S Antennas have also produced pole mounted omnidirectional circularly polarised aerials for use on structures where a cantilever position is available. These are particularly useful where a combined MF/VHF installation is needed but have also been used at several VHF-only sites.

The Alan Dick & Co system, called a 'Twin Z' for reasons which are apparent from fig.13, generates the circularly polarised signal from a pair of elements each made up of horizontal and vertical radiators arranged symmetrically around a pole to form effectively two 'arrowhead' horizontally polarised dipoles and four short vertically polarised dipoles. The 90° phase relationship between components is determined by the physical layout. A good omnidirectional hrp is produced (see fig.14) and is maintained over an adequate bandwidth.

The C&S Antenna's omnidirectional system is a recent development, currently installed at two sites. A similar concept was described in a 1947 RCA Review paper by Brown and Woodward[1] and was also proposed earlier in a 1941 article in Communications[2]. The system consists of four slanted dipoles arranged around a pole and fed by equal co-phased signals. This produces circularly polarised

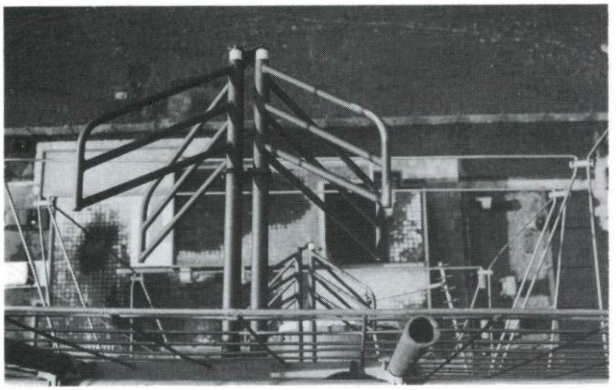

Fig.9. C&S Antennas' circularly polarised panel aerial as installed at Dover.

radiation which can be seen by considering the situation looking directly at one dipole as in fig.15. Since the wavefront from the rear element arrives at the front element with the phase reversed due to the λ/2 path difference, the resulting vector is horizontally polarised due to cancellation of the vertical components.

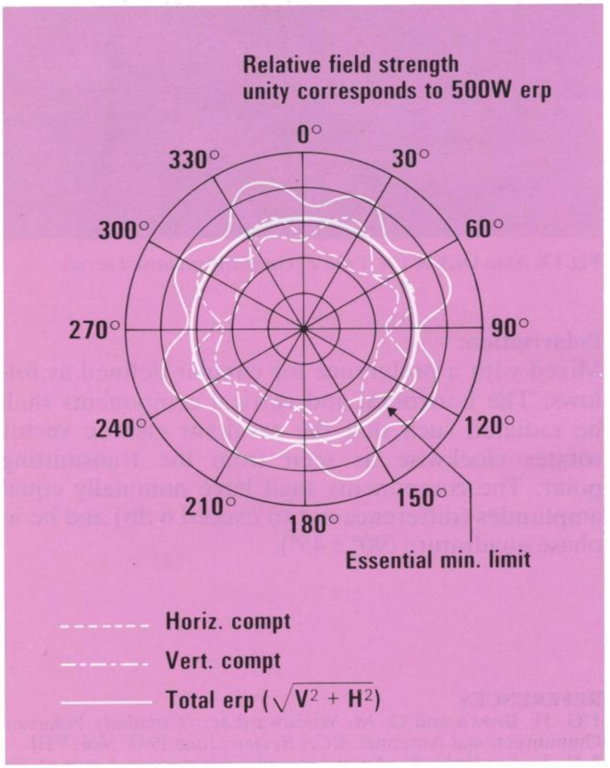

Fig.10. Radiation patterns of the Alan Dick & Co circularly polarised panel aerial shown in fig.7.

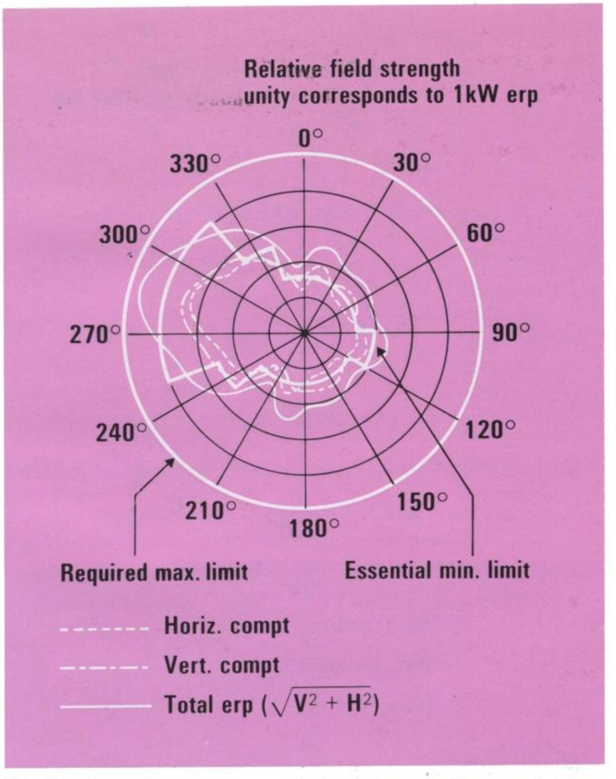

Fig.11. Radiation patterns of the Marconi Communications Systems' circularly polarised panel aerial at Brighton.

Wavefronts from the side elements add up to produce a vertically polarised resultant which has a 90° path difference to the horizontally polarised component. This therefore satisfies the requirements for circular polarisation. The actual phase relationship is affected by the presence of the pole but is quite satisfactory for IBA requirements. A very good wideband omnidirectional hrp is produced and the whole system is very compact as can be seen from figs.16 and 17.

Both systems have limitations in that they need to be top mounted and can only produce omnidirectional patterns. However, the C&S Antenna's arrangement can be modified to produce a directional hrp by horizontal displacement of adjacent tiers. The Alan Dick system has a high power handling potential because of its robust construction and also has a simpler distribution system as it only needs two inputs per tier.

Conclusion

The availability of Band II VHF radio transmitting aerial systems suitable for circular polarisation has

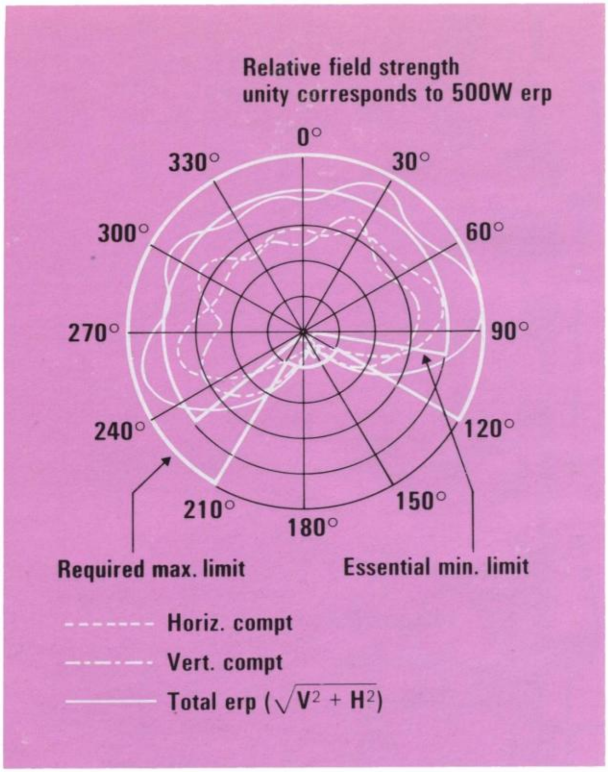

Fig.12. Radiation patterns for the C&S Antennas' circularly polarised panel aerial shown in fig.9.

improved considerably over the last few years. The shortcomings of the early wideband systems can largely be overcome, provided a support-structure with sufficient extra capacity is available. The major UK manufacturers have now come a long way towards meeting the needs of the IBA and the BBC in developing a useful range of suitable aerial systems.

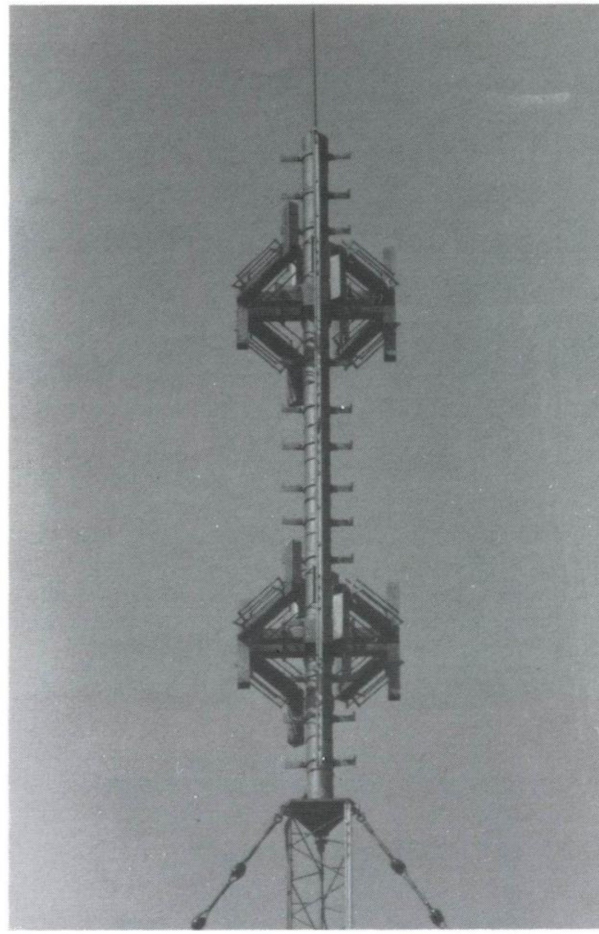

Fig.13. Alan Dick & Co 'Twin Z' circularly polarised aerial.

APPENDIX: VHF ILR Aerial Specification

The specification currently used for a typical VHF ILR aerial is:

Radiation Pattern: Directional or Omnidirectional as required.

Impedance:
50 ohms with voltage reflection coefficient not exceeding 15% over the local radio sub-bands 94.5-97.6 MHz and 100-105 MHz with a 3 db relaxation (to 21%) for half aerial working.

Polarisation:
Mixed with a preference for circular defined as follows: The horizontal and vertical components shall be radiated such that the resultant electric vector rotates clockwise as seen from the transmitting point. The components shall have nominally equal amplitudes (difference not to exceed 6 db) and be in phase quadrature (90° ± 45°).

REFERENCES
1 G. H. Brown and O. M. Woodward Jr., 'Circularly Polarised Omnidirectional Antenna', *RCA Review*, June 1947, Vol. VIII.
2 N. E. Lindenblad, 'Antennas and Transmission Lines at the Empire State Television Station', *Communications*, Vol. 21, No. 4, April, 1941.

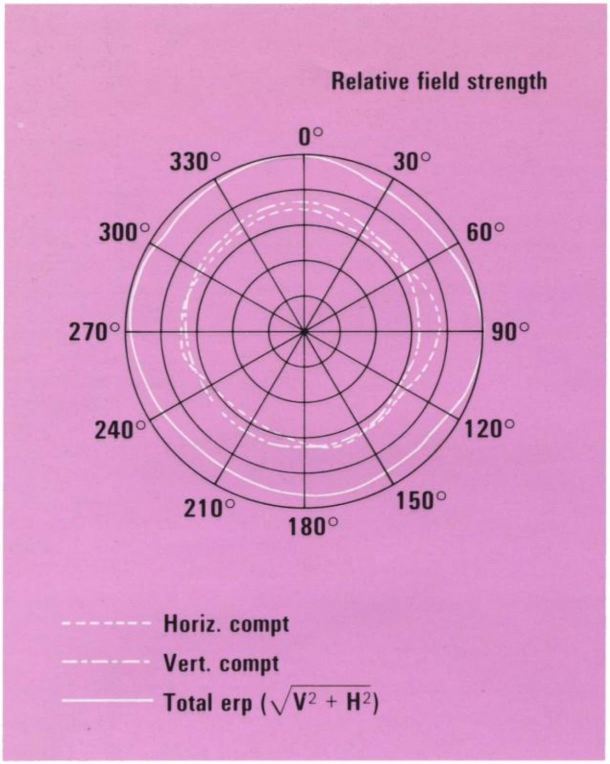

Fig.14. Radiation patterns for Alan Dick & Co 'Twin Z' circularly polarised aerial.

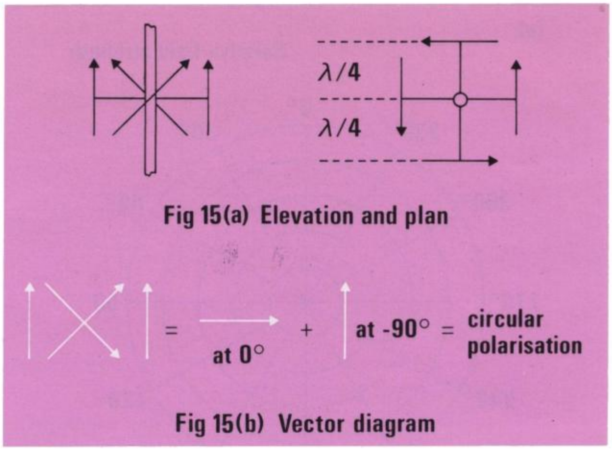

Fig.15. C&S Antennas' pole mounted omnidirectional circularly polarised aerial.
Fig.15a. Mechanical arrangement of radiating elements.
Fig.15b. Derivation of circularly polarised signal.

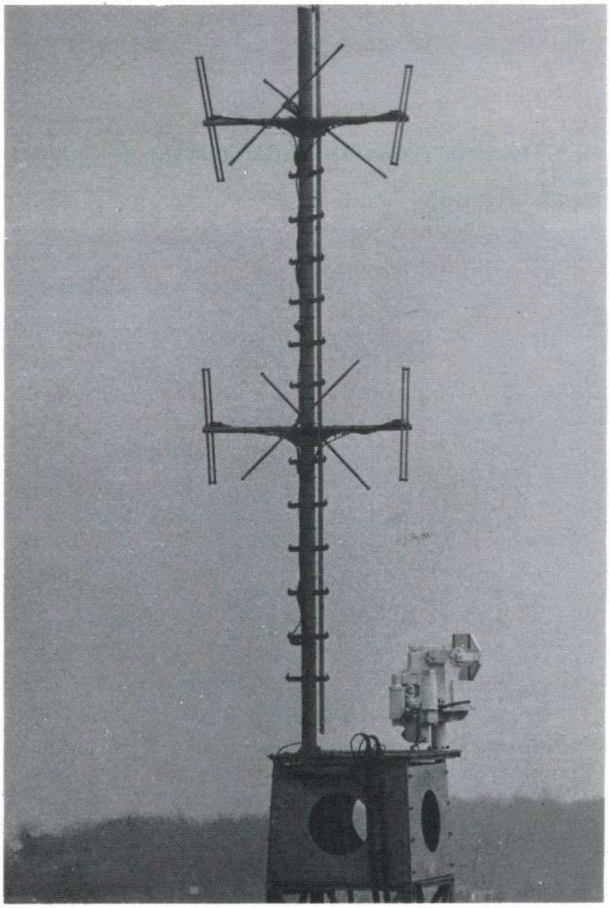

Fig.16. A C&S Antennas' pole mounted omnidirectional aerial as erected at Bluebell Hill for the East Kent ILR service.

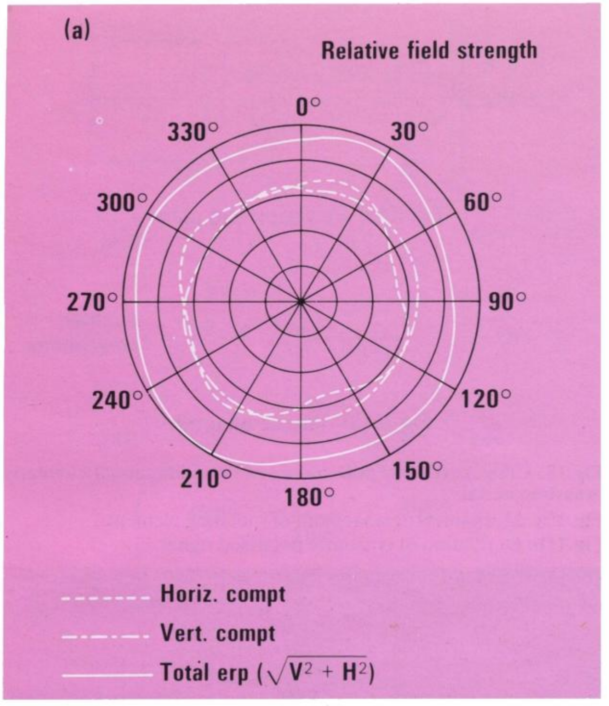

Fig.17. Radiation patterns for the C&S Antennas' pole mounted aerial.
Fig.17a. At 88 MHz.

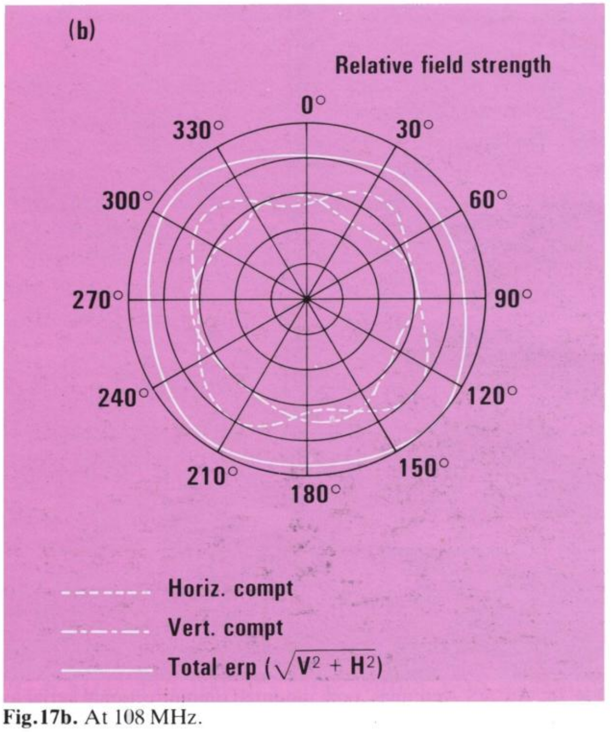

Fig.17b. At 108 MHz.

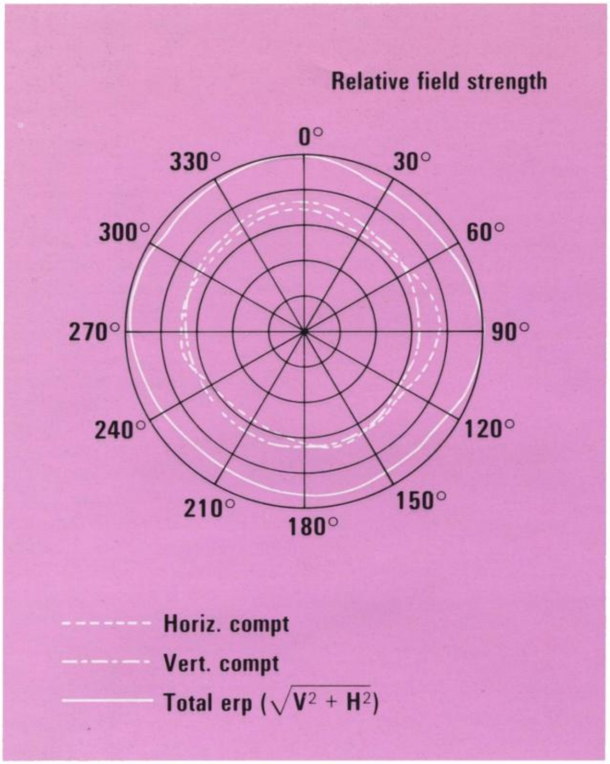

Fig.14. Radiation patterns for Alan Dick & Co 'Twin Z' circularly polarised aerial.

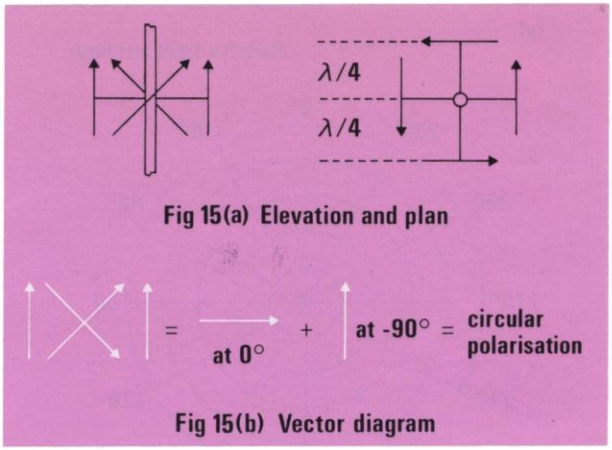

Fig.15. C&S Antennas' pole mounted omnidirectional circularly polarised aerial.
Fig.15a. Mechanical arrangement of radiating elements.
Fig.15b. Derivation of circularly polarised signal.

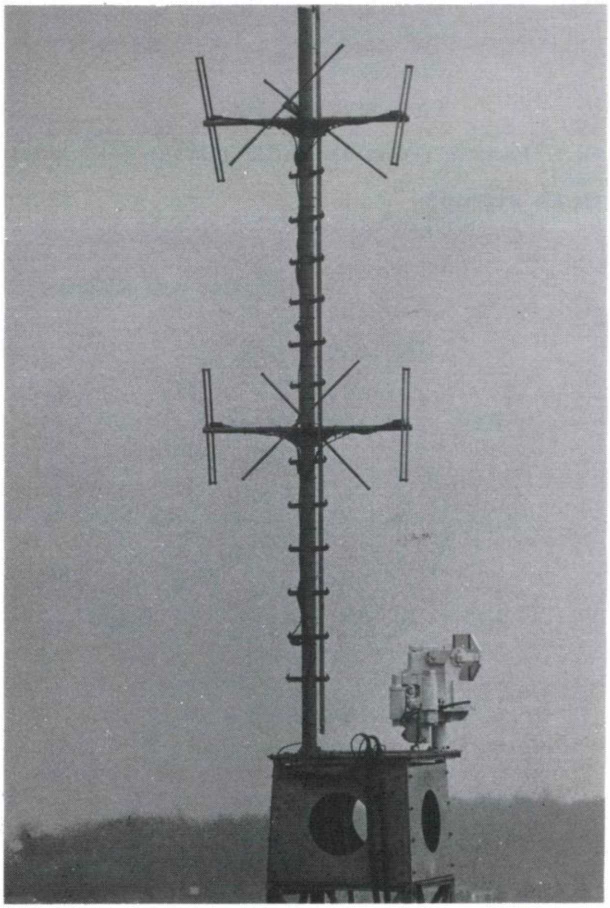

Fig.16. A C&S Antennas' pole mounted omnidirectional aerial as erected at Bluebell Hill for the East Kent ILR service.

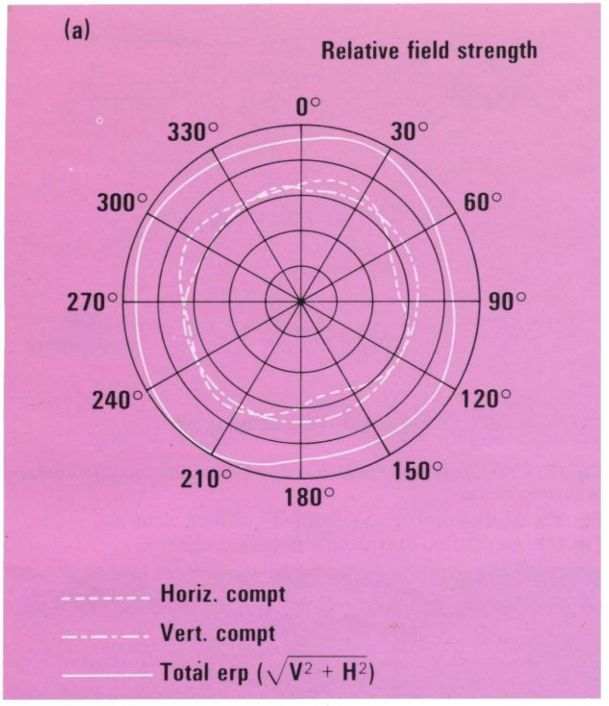

Fig.17. Radiation patterns for the C&S Antennas' pole mounted aerial.
Fig.17a. At 88 MHz.

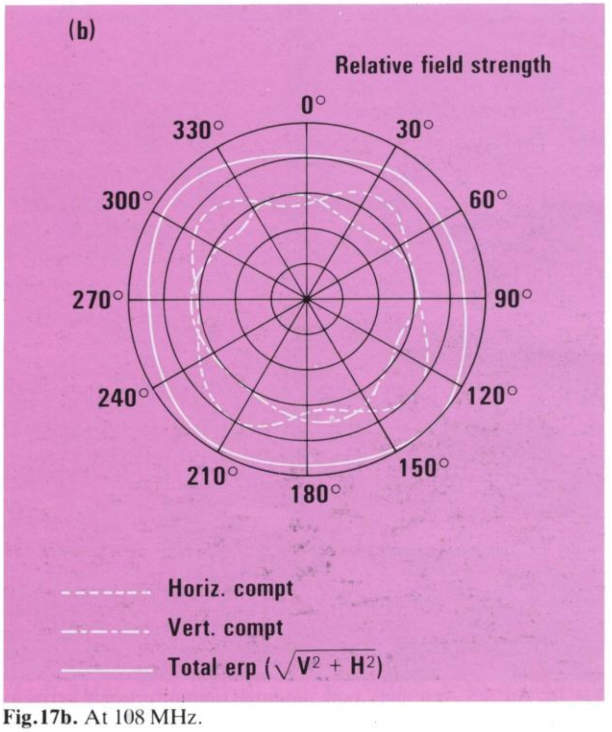

Fig.17b. At 108 MHz.

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Mise en oeuvre et entretien de materiel d'antenne IBA

par J. A. Thomas

Résumé

La responsabilité de la mise en oeuvre et de l'entretien de tout matériel et de toutes structures d'antenne au sein d'IBA a été fusionnée en un simple groupe. Un effectif total de 60, dont la majorité est basée dans les régions techniques, effectue cette tâche. La mise en oeuvre de nouveau matériel d'antenne est effectuée par une petite unité basée à l'état-major technique à Crawley Court. Les antennes en service aux stations UHF de télévision, principales et de relais, sont passées en revue ensemble avec leur équipement de surveillance de force et de protection. La mise au point d'améliorations concernant ce matériel fait l'objet d'une discussion. Des installations d'antenne et de mât de réserve sont essentielles pour les réseaux IBA de télévision et d'émission radio, et une gamme complète de matériel de réserve intermédiaire et principal est disponible dans le cas d'une défaillance catastrophique.

Structures de soutien d'antenne

par M. Lambert

Résumé

L'auteur passe en revue les nombreux facteurs qui doivent être pris en considération par l'Ingénieur de Transmission en définissant une structure de soutien convenant à un système d'antenne de transmission, et il trace les grandes lignes des méthodes actuelles d'étude et de construction de pylônes non haubannés et de mâts haubannés.

Une enquete sur la distorsion de diagramme d'antenne par des structures de rerayonnement

par N. D. Porter

Résumé

La méthode d'analyse de moments numériques peut être utilisée pour analyser les diagrammes de rayonnement d'antennes simples en présence de structures de rerayonnement. Dans de telles conditions, une hypothèse de répartition de courant uniforme ou sinusoïdale est insuffisante et une solution doit être recherchée à la vraie répartition de courant à la fois sur l'antenne et sur sa structure de soutien. L'antenne 'Dipôle-Bouclée' VHF Bande II soutenue par un poteau en porte-à-faux est analysée en détail au moyen de cette méthode avec un logiciel rédigé pour une application sur l'ordinateur Honeywell d'IBA.

Antennes MF à deux canaux

par D. Cowans

Résumé

Cet article traite les difficultés particulières rencontrées lors de la mise en oeuvre d'une antenne à fréquence moyenne ayant plus d'un canal. Il se concentre sur des systèmes non-directionnels et démontre les voies dans lesquelles des performances optimales peuvent être obtenues, y compris une étude. Les performances de filtre sont traitées avec quelque détail.

Antennes MF directionnelles avec reflecteurs à fil incline

par M. Zapitis

Résumé

Le système d'antenne décrit comprend un mât haubanné soutenant un réflecteur passif à fil incliné. Lorsque le mât est alimenté à son base et le réflecteur est accordé passivement, ce système produit un diagramme de rayonnement directionnel et horizontal en cardioïde. L'entreprise IBA a mis en service avec succès dix stations MF semblables depuis 1975. Les rapports avant/arrière s'étendent entre 5dB et 17dB avec un gain de faisceau principale mesuré et allant jusqu'à +2,5dB.

L'avantage de ce système sur une antenne à deux mâts, dont un passif, se traduit par une économie en coûts d'environ 20%, réalisée en se passant du second mât.

Adaptation d'impedance à multi-canaux à UHF/VHF

par C. Jacob

Résumé

La condition de la Loi sur la télévision britannique de 1963, de situer ensemble les émetteurs IBA et BBC pour le réseau de télévision national UHF a conduit à l'utilisation, dans la plupart des stations, d'une antenne simple d'émission capable de rayonner les quatre services simultanément. Des nouvelles techniques d'adaptation de ligne d'émission ont été élaborées pour surmonter les problèmes de satisfaire les limitations rigoureuses ROS nécessaires dans de tels systèmes, afin de réduire le rayonnement d'images retardées, et quelques unes de ces techniques, qui ont également une application générale, sont développées ci-dessous. Cela est suivi par une revue sommaire des méthodes d'étalement de bande et l'article se termine par une discussion sur quelques unes des natures pratiques associées au mesurage et à la correction d'impédance à RF.

ROS = Rapport d'Ondes Stationnaires.

Materiel de mixage de canaux Rotamode

par E. T. Ford

Résumé

A cette date (1985), il y a 160 Unités Rotamode de Marconi de différents types en service dans 49 stations principales UHF. Une grande partie consiste en des mixeurs de canaux, montés par paires assorties afin de permettre à la quatrième chaîne de télévision britannique d'être alimentée dans les antennes d'émission existantes à multicanaux aux stations principales IBA et BBC. Dans presque tous les cas, ces mixeurs comportent deux canaux seulement, mais dans un aménagement unique à Bilsdale, le matériel devait être monté à un emplacement où il portait tous les quatre services.

Le mixeur de canaux UHF Rotamode comprend une paire de cavités jumelées dans un guide d'ondes cylindrique, contenant des boucles de couplage de lignes d'entrée et de sortie. Son mode de fonctionnement est décrit d'une façon simplifiée et les performances mesurées du matériel de Bilsdale sont présentées.

Antennes d'emission radio VHF

par D. R. Brian

Résumé

Le besoin d'assurer des antennes d'émission VHF polyvalentes et économiques pour la Radio Locale Indépendante (ILR) a posé à l'entreprise IBA et à l'industrie de fabrication d'antennes quelques problèmes exceptionnels. Les variations dans les conditions électriques et mécaniques, à la fois générales et spécifiques, ont conduit à un perfectionnement considérable dans les récentes années, et cet article contient l'historique en arrière-plan des problèmes et trace les grandes lignes de quelques unes des solutions. Les premiers systèmes comportaient des éléments à canal simple mais un perfectionnement supplémentaire pour couvrir le nouveau besoin de large bande a conduit aux problèmes de diagramme de rayonnement. L'inclusion d'une polarisation en pente comme une option a augmenté le choix d'antennes disponibles, mais des récents perfectionnements à la fois dans les systèmes polarisés de façon circulaire directionnels et omnidirectionnels apportés par les principaux fabricants britanniques ont élargi davantage le domaine des systèmes d'antenne d'emmision radio VHF.

Die Beistellung und Wartung von IBA-Antennengerät

von J. A. Thomas

Abriß

Die Verantwortlichkeit für die Beistellung und Wartung aller Antennengeräte und -bauwerke der IBA ist in eine einzige Gruppe zusammengefaßt worden. Rund 60 Mitarbeiter, deren Mehrzahl in den technischen Regionen basiert ist, sind in diesem Aufgabenbereich beschäftigt. Die Beistellung von neuem Antennengerät ist Sache einer kleinen Einheit, die im technischen Hauptquartier in Crawley Court ansässig ist.

Es wird eine Übersicht der in den UHF-Fernseh-Haupt- und Relaissendern gebräuchlichen Antennen gegeben und deren Leistung, Überwachungs- und Schutzausrüstungen angeführt, und es wird die Entwicklung von Verbesserungen an diesen Ausrüstungen erörtert.

Für die Fernseh- und Rundfunknetze der IBA sind Reserveantennen und -maste von wesentlicher Bedeutung und für den Fall eines katastrophalen Ausfalls steht ein voller Bereich an Zwischen- und Hauptreservegerät zur Verfügung.

Antennentragbauwerke

von M. Lambert

Abriß

Der Verfasser betrachtet die vielen Einflußfaktoren, die vom Rundfunk-Ingenieur bei der Definition eines geeigneten Tragbauwerkes für ein Rundfunkantennensystem berücksichtigt werden müssen, und er beschreibt die jetzigen Konstruktions- und Baumethoden von freistehenden und abgespannten Masten.

Eine Untersuchung der Antennenbildverzerrung durch Zwischensende-Bauwerke

von N. D. Porter

Abriß

Mit der Untersuchungsmethode der numerischen Momente kann man die Strahlungsbilder von einfachen Antennen beim Vorhandensein von Zwischensendebauwerken untersuchen. Unter solchen Bedingungen ist die Annahme einer gleichförmigen oder sinusförmigen Stromverteilung nicht angebracht, und es muß eine Lösung für die wahre Stromverteilung an der Antenne wie auch an ihrem Tragbauwerk gefunden werden. Es wird die auf einem freitragenden Mast gestützte VHF Band II "Loop-Dipole"-Anordnung mit dieser Methode im einzelnen untersucht, und zwar mit Hilfe einer Software, die für Gebrauch mit dem Honeywell Computer der IBA erstellt worden ist.

Zweikanal-MF-Antennen

von D. Cowans

Abriß

Dieser Artikel behandelt die besonderen Schwierigkeiten, die beim Belasten einer MF-Antenne mit mehr als einem Kanal entstehen. Er konzentriert sich auf Rundstrahlsysteme und veranschaulicht die Möglichkeiten, mit welchen sich optimale Leistung erzielen lassen und enthält auch eine Fallstudie. Filterleistungen werden etwas ausführlicher behandelt.

MF-Richtantennen mit Schrägen Drahtreflektoren

von M. Zapitis

Abriß

Das beschriebene Antennensystem besteht aus einem abgespannten Mast, der einen passiv abgestimmten schrägen Drahtreflektor trägt. Wird der Mast an seiner Basis eingespeist und der Reflektor ist passiv abgestimmt, erzeugt dieses System ein herzförmiges, horizontales Strahlungsrichtdiagramm.

Seit 1975 hat die IBA zehn solcher MF-Sender erfolgreich in Betrieb gesetzt. Die Vor-Rück-Verhältnisse erstrecken sich über einen Bereich von 5 dB bis 17 dB mit einer gemessenen Hauptsendestrahlverstärkung von bis zu +2,5 dB.

Verglichen mit einer passiv abgestimmten Zweimast-Anordnung hat dieses System den Vorteil, daß eine Kostenersparnis von etwa 20% durch den Wegfall des zweiten Mastes eintritt.

Mehrkanal-Impedanzanpassung für UHF/VHF

von C. Jacob

Abriß

Die im britischen Fernsehgesetz 1963 enthaltene Vorschrift, IBA- und BBC-Sender für das nationale UHF-Fernsehnetz gemeinschaftlich anzuordnen, führte bei den meisten Sendestationen zu dem Gebrauch einer einzigen Sendeantenne, die imstande ist, alle vier Sendedienste gleichzeitig auszustrahlen. Um die Probleme zu lösen, die sich mit der Erfüllung der für solche Systeme erforderlichen anspruchsvollen VSWR-Begrenzungen einstellen, sind neuartige Sendeleitungsanpaßverfahren entwickelt worden, welche die Ausstrahlung von verzögerten Bildern herabsetzen und einige dieser Verfahren, die auch eine allgemeine Anwendung haben, werden im Folgenden entwickelt. Dem schließt sich eine Übersicht der Breitbandverfahren an und der Aufsatz endet mit einer Erörterung einiger der praktischen Aspekte im Zusammenhang mit der Impedanzmessung und -korrektur bei Radiofrequenz.

Rotamode-Kanalmischgerät

von E. T. Ford

Abriß

Es stehen 160 Marconi Rotamode-Geräte verschiedener Arten in 49 UHF-Hauptsendern zum heutigen Zeitpunkt (1985) im Einsatz. Ein großer Teil dieser sind in abgestimmten Paaren installierte Kanalmischgeräte, die es ermöglichen, den vierten britischen Fernsehkanal in bestehende Mehrkanal-Sendeantennen in den Hauptsendern der IBA und des BBC einzuspeisen. Diese Mischgeräte tragen in fast allen Fällen nur zwei Kanäle, in einer einzigartigen Anordnung in Bilsdale mußte die Ausrüstung aber an einem Ort installiert werden, wo sie alle vier Sendedienste trägt.

Das Rotamode-UHF-Kanalmischgerät besteht aus einem Paar gekoppelter Hohlräume in einem zylindrischen Hohlleiter, der Koppelschleifen von Eingangs- und Ausgangsstreifenleitungen enthält. Seine Betriebsweise wird in einfacher Art und Weise erläutert und es wird die gemessene Leistung der Bilsdale-Ausrüstung vorgetragen.

VHF-Funkantennen

von D. R. Brian

Abriß

Das Erfordernis, dem Unabhängigen Ortsrundfunk (Independent Local Radio = ILR) anpassungsfähige und wirtschaftliche VHF-Funkantennen zu verschaffen, stellte die IBA und die Antennenindustrie vor eine Reihe von einzigartigen Problemen. Die Abweichungen in den allgemeinen wie auch spezifischen elektrischen und maschinenbaulichen Erfordernissen führten zu ganz beträchtlichen Entwicklungen in den letzten Jahren und dieser Artikel beschreibt den Werdegang der dahinterstehenden Probleme und einige der Lösungen. Ehere Systeme wirkten mit Einkanalelementen, doch führte die weitere Entwicklung zur Deckung des Erfordernisses auf ein neues Breitband zu Problemen in der Strahlungscharakteristik. Mit der Schrägpolarisierung als zusätzliche Option hat sich die Auswahl an erhältlichen Antennen vergrößert, in letzter Zeit gemachte Entwicklungen mit kreispolarisierten Richt- und Rundstrahlsystemen bei bedeutenden britischen Herstellern haben aber die Möglichkeiten für VHF-Funkantennensysteme noch mehr erweitert.

Provisión y mantenimiento de equipo de antenas IBA

por J. A. Thomas

Resumen

La provisión y mantenimiento de todo el equipo y estructuras de antenas en la IBA se ha puesto a cargo de un solo grupo. Unas sesenta personas, pertenecientes la mayoría a la sección de ingeniería, llevan a cabo esta tarea. La provisión de nuevo equipo de antenas se realiza por una pequeña unidad basada en la oficina central de ingeniería en Crawley Court.

Aquí se pasa revista a las antenas utilizadas en las estaciones de relé y principales de televisión de UHF, junto con su equipo de protección y comprobación de potencia. Se discute también los desarrollos para la mejora de este equipo.

Es fundamental disponer de antenas y mástiles de reserva para las redes de radiotransmisión y televisión de IBA, disponiéndose de una serie completa de equipo de reserva principal y provisional, para caso de fallo catastrófico.

Estructuras para soporte de antenas

por M. Lambert

Resumen

El autor revisa los muchos factores que deben tenerse en cuenta por el ingeniero al definir una estructura soporte adecuada para un sistema de antena de radiodifusión, e indica los métodos actuales de diseño y construcción de torres autosoportadas y mástiles arriostrados.

Investigación sobre distorsión de diagramas de radiación por estructuras rerradiadoras

por N. D. Porter

Resumen

El método analítico de momentos numéricos puede utilizarse para analizar los diagramas de radiación de antenas sencillas en presencia de estructuras rerradiadoras. En tales condiciones no es adecuado suponer una distribución de corriente uniforme o senoidal, debiendo encontrarse una solución para la distribución de corriente verdadera, tanto en la antena como en su estructura soporte. El conjunto de "dipolo de cuadro" banda II de UHF, soportado sobre un poste voladizo, es analizado en detalle empleando este método, con software escrito para ser usado en la computadora Honeywell de IBA.

Antenas de frecuencia media de canal doble

por D. Cowans

Resumen

Este artículo trata de las dificultades especiales encontradas para alimentar una antena de frecuencia media con más de un canal. Se concentra en sistemas no direccionales y demuestra el modo de obtener un rendimiento óptimo, incluyendo el estudio de un caso particular. Se considera con detalle el rendimiento de los filtros.

Antenas direccionales de frecuencia media con reflectores de conductor inclinado

por M. Zapitis

Resumen

El sistema de antena descrito consta de un mástil arriostrado que soporta un reflector de conductor inclinado sintonizado parasíticamente. Cuando se alimenta el mástil en su base y el reflector está sintonizado pasivamente, este sistema produce un diagrama de radiación horizontal de cardiode direccional.

La IBA ha instalado diez de dichas estaciones de frecuencia media con todo éxito desde 1975. La eficacia direccional oscila entre 5dB y 17dB, con una ganancia de haz principal medida de hasta +2,5dB. La ventaja de este sistema sobre un conjunto sintonizado parasíticamente de dos mástiles, es el ahorro del 20 p.c. aproximadamente obtenido al utilizar sólo un mástil.

Adaptación de impedancia multicanal en VHF/UHF

por C. Jacob

Resumen

El requerimiento de la ley de televisión del Reino Unido del año 1963, por el que los transmisores de la IBA y BBC para la red de televisión nacional de UHF deben tener un emplazamiento común, ha conducido al empleo, en la mayoría de las estaciones, de una antena transmisora única capaz de radiar simultáneamente los cuatro servicios. Se han desarrollado nuevas técnicas de adaptación de líneas de transmisión, para poder satisfacer las severas limitaciones de ROE necesarias en dichos sistemas, con el fin de reducir la radiación de imágenes retardadas, describiéndose aquí algunas de ellas, que tienen también aplicación general. A continuación se hace una revisión general de los métodos de ensanche de banda, concluyendo con una discusión sobre algunos de los casos prácticos asociados con las medidas de impedancia y la corrección en RF.

Equipo de combinación de canales Rotamode

por E. T. Ford

Resumen

Actualmente (1985) hay 160 Unidades Rotamode Marconi de diversos tipos en servicio en 49 estaciones principales de UHF. Una gran proporción son combinadores de canales, instalados en pares acoplados para permitir que el cuarto canal de televisión del Reino Unido sea alimentado a las antenas de transmisión multicanal existentes en las estaciones principales de la IBA y BBC. En casi todos los casos estos combinadores llevan sólo dos canales, pero en un dispositivo especial en Bilsdale el equipo tuvo que ser instalado en un lugar donde lleva los cuatro servicios.

El combinador de canales de UHF Rotamode comprende un par de cavidades acopladas en guiaondas cilíndrico, conteniendo bucles de acoplamiento de línea de cinta de entrada y salida. Se describe su método de funcionamiento de un modo simplificado, presentándose el rendimiento obtenido con el equipo de Bilsdale.

Antenas de radiotransmisión de VHF

por D. R. Brian

Resumen

La necesidad de disponer de antenas de transmisión de VHF versátiles y económicas para Radio Local Independiente (ILR) ha planteado a IBA y a la industria de fabricación de antenas algunos problemas especiales. Las variaciones de requisitos generales y específicos, tanto eléctricos como mecánicos, ha conducido a un considerable desarrollo en los últimos años, describiendo este artículo la base de estos problemas e indicando algunas de las soluciones adoptadas. Los primeros sistemas empleaban elementos de canal único, pero el desarrollo posterior para cumplir un nuevo requisito de banda ancha condujo a problemas de diagrama de radiación. La inclusión de polarización real como opción, aumentó el surtido de antenas disponible, pero recientes desarrollos en sistemas polarizados circularmente tanto direccionales como omnidireccionales, por los principales fabricantes del Reino Unido, han ampliado más el ámbito de los sistemas de antenas de radiotransmisión para VHF.

IBA TECHNICAL REVIEW

1 Measurement and Control*
2 Technical Reference Book (3rd Edition)*
3 Digital Television*
4 Television Transmitting Stations*
5 Independent Local Radio*
6 Transmitter Operation and Maintenance*
7 Service Planning and Propagation*
8 Digital Video Processing – DICE*
9 Digital Television Developments*
10 A Broadcasting Engineer's Vade Mecum*
11 Satellites for Broadcasting*
12 Techniques for Digital Television
13 Standards for Television and Local Radio Stations*
14 Latest Developments in Sound Broadcasting*
15 Microelectronics in Broadcast Engineering
16 Digital Coding Standards
17 Developments in Radio-frequency Techniques
18 Standards for Satellite Broadcasting
19 Technical Training in Independent Broadcasting
20 Developments in Teletext
21 Compatible Higher-Definition Television
22 Light and Colour Principles

* *Out of Print*

The Independent Broadcasting System
The Independent Broadcasting Authority (IBA) is the central body responsible for the provision of Independent Television (ITV, including TV-am, and Channel 4) and Independent Local Radio (ILR) services in the United Kingdom. The IBA selects and appoints the programme companies; supervises the programme planning; controls the advertising; and builds, owns and operates the transmitters. Independent Broadcasting is completely self-supporting, financed by the sale of spot advertising time in the companies' own areas.

More information is contained in *Television & Radio 1986*, the IBA's guide to Independent Broadcasting (£3.90 from bookshops).

Printed by J Looker Printers Ltd, 50 Willis Way, Poole, Dorset.

IBA INDEPENDENT BROADCASTING AUTHORITY